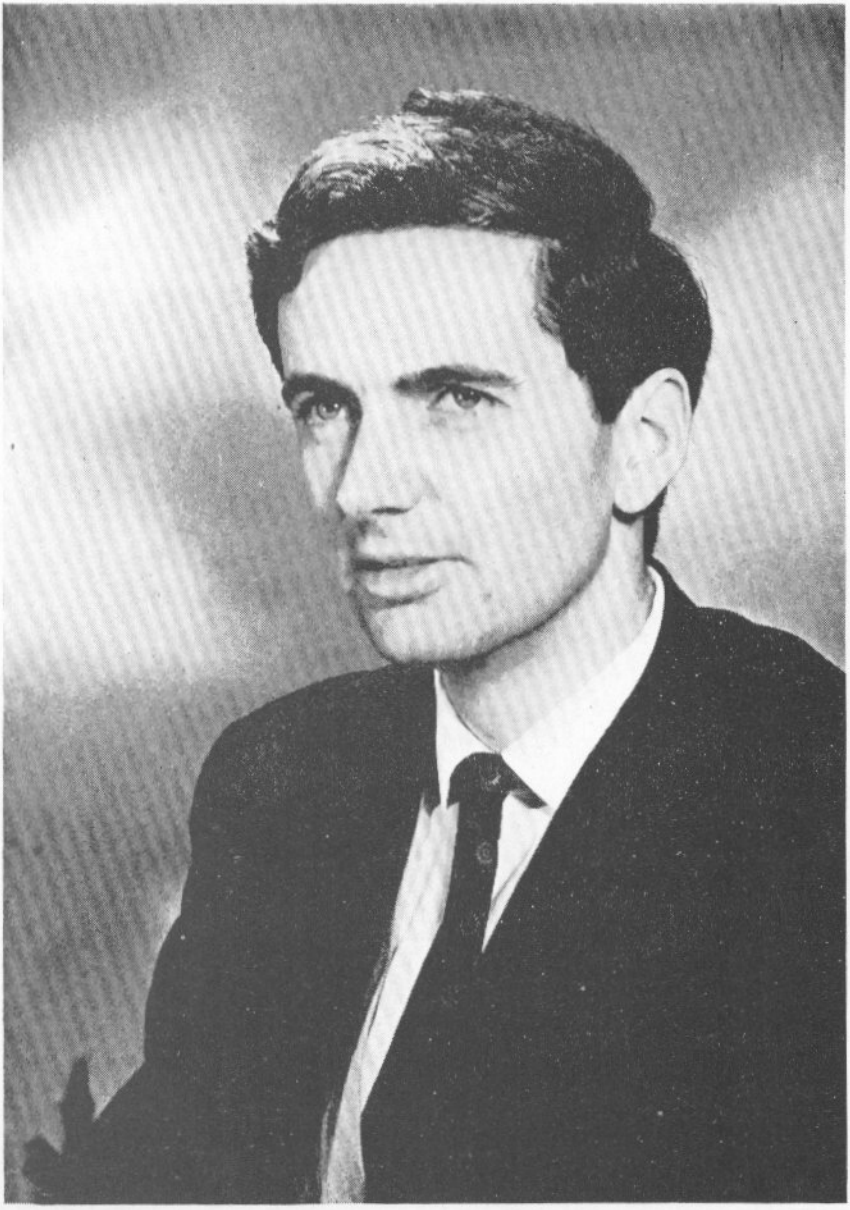
theory and practice of

MODEL RADIO CONTROL

third edition with latest circuitry

development of systems
proportional explained
digital system design-
practical examples and
full circuit details

Paul Newell



The author

THEORY AND PRACTICE OF MODEL RADIO CONTROL

by Paul Newell, B.Sc.

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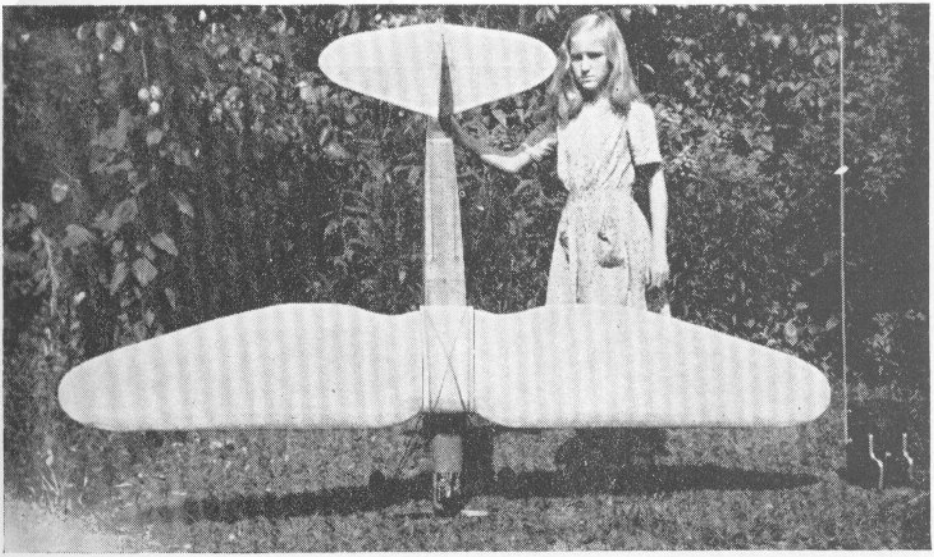
INTRODUCTION

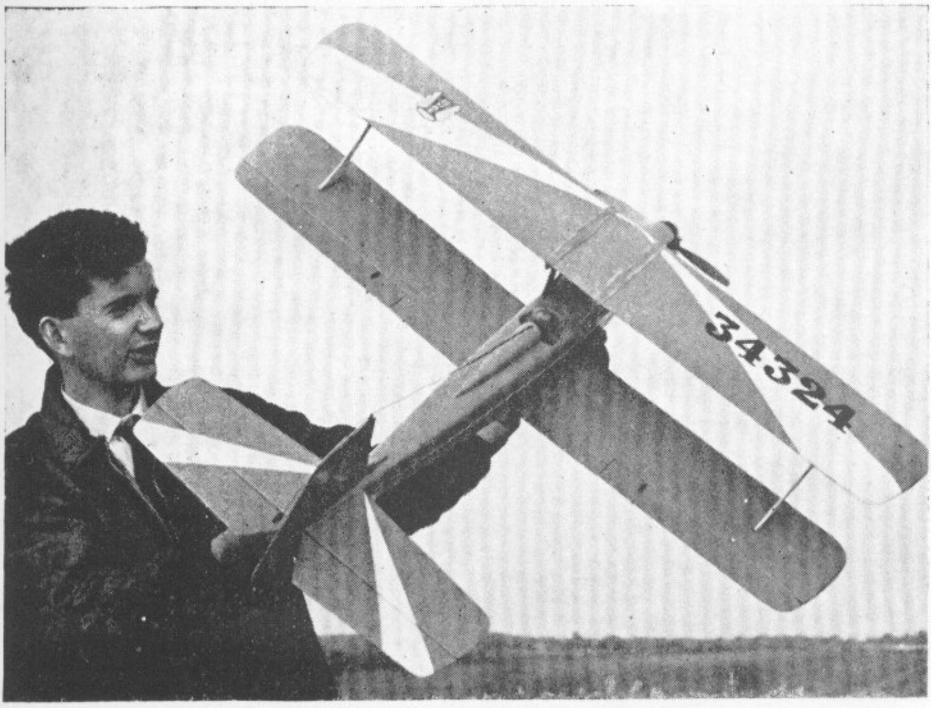
THE purpose of this book is to provide modellers with an insight into what makes proportional equipment operate. We have attempted to build up the explanation of what is a most sophisticated system, in such a manner that no knowledge of electronics is required, and we hope that the reader, after studying the initial chapters, will have gained sufficient knowledge to appreciate the circuit operational descriptions in the later chapters.

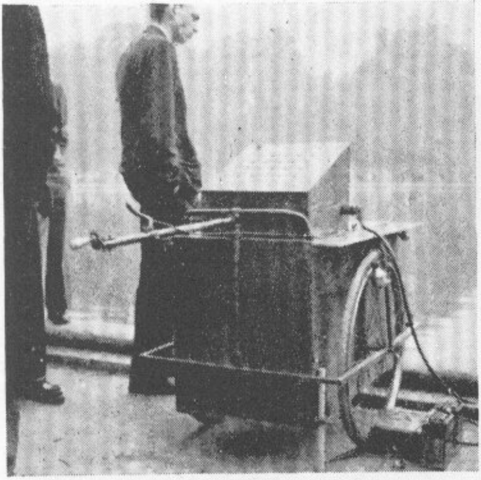
Those readers with some electronics knowledge should not find that this book is outside their range of interest since, to our knowledge, all aspects of proportional equipment design have never before been combined in one volume.

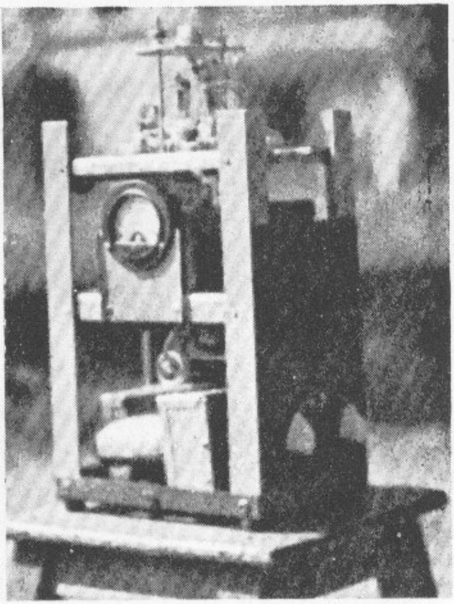
In order to provide a full description, it has been necessary to include example circuits of a typical system. The author's "Microtrol" five channel system has been used for this purpose since, over a period of time, this has proved a repeatable and reliable piece of equipment. Constructional notes are included in the appendices, but it must be stressed that the purpose of this book is *not* that of a constructional handbook. Those readers with adequate knowledge and experience will find the notes more than sufficient. A reader contemplating home construction, but having doubts about his ability to produce a satisfactory system from the information given, is best advised to consider one of the many excellent commercial kitted systems available. *It is regretted that neither the publishers nor the author can enter into any correspondence concerning the Microtrol system.*

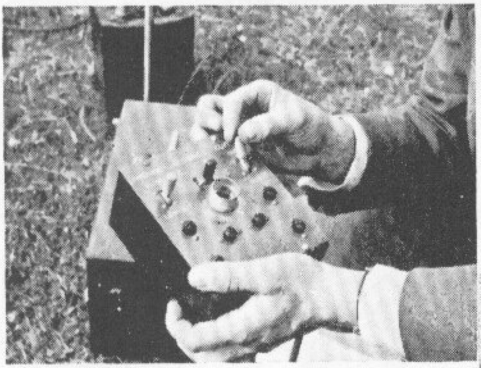


Compare this 8lb. single channel model of 1948, built by pioneer George Honnest-Redlich, with a contemporary 39in. span biplane—R.M.'s *Pasendena Special*—using four function proportional control

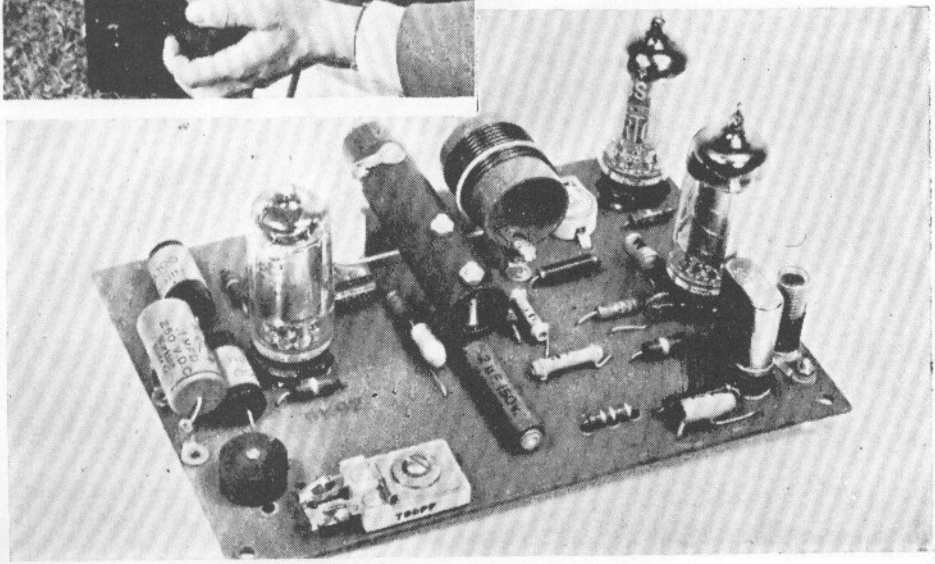


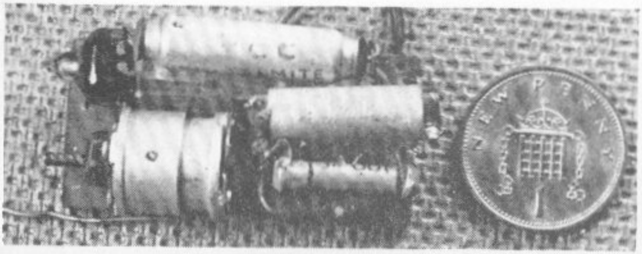


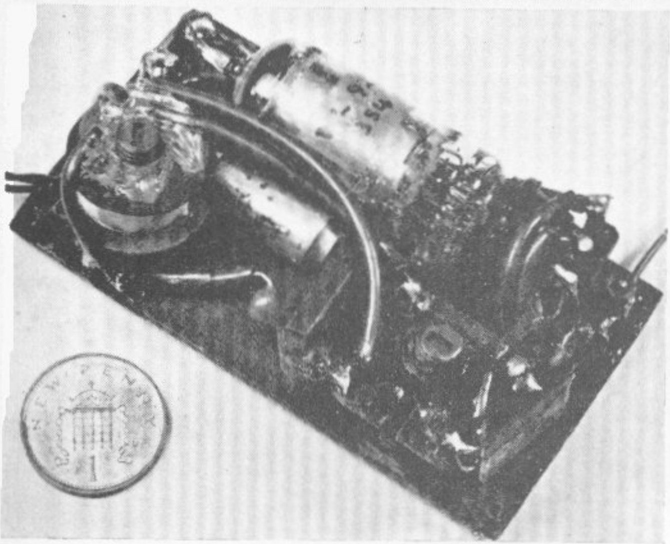


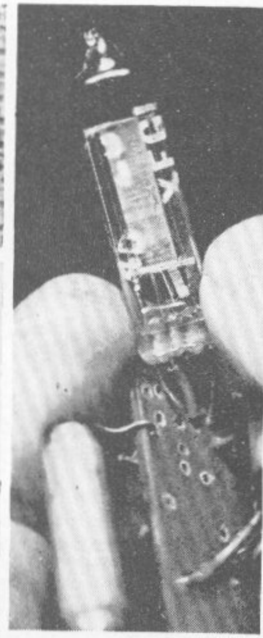


Early transmitters were not easily portable, as this "ice cream cart" single channel uniselector type, at a French boat contest over 25 years ago, shows. Above: another single channel transmitter with option of tone or carrier. Left: ground based transmitters often featured hand-held control-boxes like this example for a six tone (three function) outfit. Below: a hand-held crystal controlled single channel tone transmitter—note pot. core for tone stability

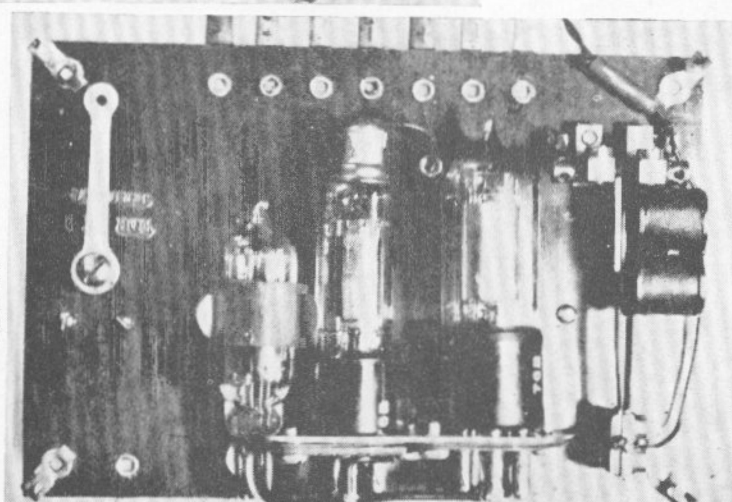








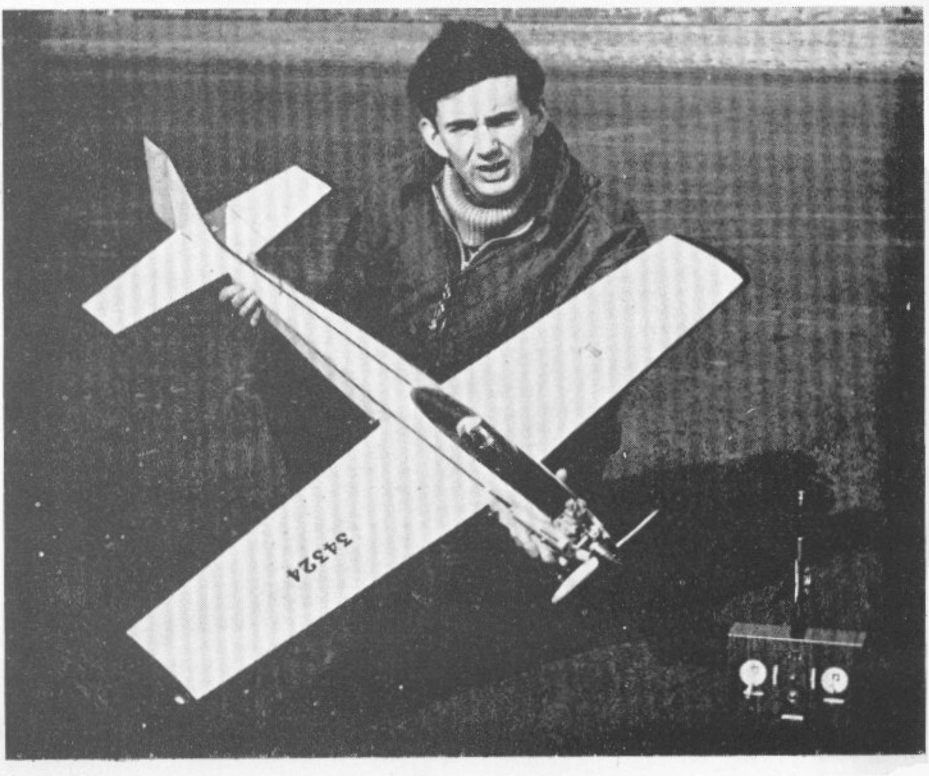
Top: this early super-regen receiver, though tiny itself, required 45v. H.T. and 1.5v. batteries for its single valve—the XFG-1—shown above. However, they did make the 36in. span model a reality



More consistent results were obtained with the larger units (above and right) of the single or 3-valve type. The former incorporates a quench coil

Sqd.-Ldr. Eric Cable with a stable single channel model owing much to free-flight design thought—*circa* 1948

Fast and aerobatic miniature "full-house" models like this are possible only with modern commercial proportional systems, or those home-builts such as are described at the end of this book



CHAPTER 1

THE EARLY DAYS

RADIO control, as we know it today, has progressed rapidly from its earliest beginnings. The majority of development has occurred since the end of the Second World War, but the concept of controlling operations by radio dates from the time a mechanical device was first employed to record Morse code.

One of the first demonstrations of radio control was at an exhibition in Paris in 1905, when Professor Branly demonstrated the firing of guns and the operation of machinery, controlled from a distance by a spark transmitter and receiver. (This type of transmission has long been superseded, and is prohibited by law due to the high level of interference it causes to other transmissions.) From this time onwards radio control progressed rapidly under military incentives so that, by the time of the Second World War, it had found application in many areas, the most well known perhaps being the German V2 rockets. Peaceful applications were not entirely ignored and, in France, control of unmanned lighthouses was performed by radio.

It is perhaps interesting that the idea of applying radio control to models appears to have come, initially, not from modellers, but rather from radio engineers. American radio periodicals first published articles on the subject in 1934. Almost universally, transmitting regulations prevented other than licensed amateurs from experimenting and, consequently, progress was slow. In France several teams appeared, comprising both modellers and radio amateurs, with the result that many forward progressive steps were made very rapidly since each could apply themselves to their respective interests.

The equipment in these early days employed the simplest principles out of pure necessity. Since the transistor was yet to come, the use of valves meant that models had to carry both high and low voltage batteries. Circuits were designed drawing minimal current to economise on high voltage batteries, and relays were used to convert relatively small current changes into a form suitable for operating electromechanical actuator devices. Since, initially, none of the components was purpose made for model control, the equipment was heavy, bulky and fragile. Whilst this was tolerable for the transmitter, it posed great problems at the receiving end, particularly for aircraft use.

I have said that the simplest principles were used, and the degree of control available was similar to that available today from single channel equipment. There was, however, one big difference, and this was in the method of transmitting information. Nowadays, single channel transmitters radiate radio frequency power continuously, and control information is sent by switching on and off a tone, that is superimposed on the radiation. The receiver detects the presence or absence of the tone in the R.F. signal, and produces an output that follows the tone switching. The early equipment did not employ a tone, but instead switched the radio frequency signal on and off. The receiver then had to detect the presence or absence of R.F. radiation. Since the strength of an R.F. signal falls off rapidly as the distance from the transmitter is increased, it becomes more difficult to interpret correctly whether the R.F.

signal is present or not. The early receivers working on this principle, therefore suffered from a varying output with signal strength. The circuits consequently had to be set up so that a relay could be operated on the minimum signal level and, in this condition, they became susceptible to changes in battery voltage. In order to get reliable performance, a complete range checkout procedure had to be performed at frequent intervals.

A very significant development took place in 1938. This was the introduction of a thyratron valve, especially designed for radio control applications. This type of valve has a trigger characteristic and so minimises the output signal variation with changing signal strengths. Unfortunately, it suffers from the disadvantage of a comparatively short useful life. A British manufactured miniaturised thyratron became available in the early 1950s, and Geoff Pike of Nottingham established a world duration record in 1954, using his own design receiver employing one of these valves.

Commercial equipment at this time was almost universally single channel, but some individuals were beginning to experiment with tuned reeds to obtain multiple control functions. Prior to this, multiple control was obtained by mechanically decoding keyed sequences. The keying was either performed manually or, in the case of slow moving models, a telephone dial was employed with a uniselector for decoding. (A uniselector is a form of rotary switch, which is stepped through its positions by pulsing an electromagnet.) These early methods obviously had the great disadvantage of a period of time elapsing whilst the mechanical decoding occurred, after the signal had been sent. The concept of tuned reeds was brought about by adopting the principle of modulating the transmitted carrier signal with tones in the audio frequency range. These tones, in transmission theory, are called sub-carriers, and the purpose of the reed bank is to filter out these sub-carriers and provide separate outputs for each control channel. The tones are adjusted to coincide with the resonant frequency of the reed and, by applying the tone to an electromagnet adjacent to the reed, it is caused to vibrate at a frequency defined by its mechanical properties. Contacts are positioned so that, as the reed vibrates, it touches and completes a circuit and operates the control actuators. The early experimenters can hardly have visualised the manner in which this form of control was to progress, either from the miniaturisation of reed banks, or the elimination of relays following the introduction of transistors.

Transistors first made their debut on the radio control scene around 1960. At this time their cost was prohibitive and they were susceptible to temperature variations. They did, however, find application in single channel equipment. By this time the standard form of single channel receiver was a valve super-regenerative detector followed by a valve amplifier stage, and the transmitted signal employed a tone to convey the control information. Transistors were introduced to replace the amplifier stage, resulting in the elimination of one valve and so reducing vulnerability to damage in aircraft applications. Gradually, as prices came down, transistors were applied to other parts of the circuits, and, by the mid 1960s, virtually all commercial equipment was transistorised throughout.

It is perhaps not unfair to say that the real incentive for reliable multi-channel control was from the aircraft field. In the space of about five years the standard of flying had reached a level that was almost unseen around 1960. This was mainly achieved by a tremendous improvement in reliability, one American manufacturer being particularly successful in this respect. The concentration on the development of reed banks over the preceeding ten years, had resulted in a channel selection device of excellent stability and reliability, but the oscillators in the transmitter for producing the tones left much to be desired. Gradually, as transistor circuit techniques improved, this problem was overcome and, since R.F. power was adequate, although perhaps somewhat inefficiently produced, the communication link presented only a few minor problems.

Such was not the case with servos, and none of those readily available offered really consistent reliability, until the first

Bonner servo was introduced. Improvements on this, and the design of a special motor, led to the Bonner Duramite servo, which was to lead the field for several years.

With this new standard of reliability, model aircraft design progressed from stable high wing types, such as the *Radio Queen*, *Falcon*, and others, to low wing aerobatic craft that required continual correction to maintain the required flight path. With the introduction of aileron control and improvements in engine power and throttling, flying performance changed almost overnight.

Transistors had improved in performance, and types capable of switching large currents became available at a reasonable price. This allowed the introduction of solid state servo amplifiers, in place of the pairs of relays required for each servo and, apart from the obvious advantage of the saving in size, reliability was further improved by a reduction in the current that the reed bank contacts were required to carry. The reed bank inself never had a chance to become outdated, because the development of tuned filters, which might have caused this, was overshadowed by the coming of proportional.

Obviously any development history contains many overlapping factors and radio control is no exception. Thus the early experimental work on proportional began many years before reed control had reached the advanced stage just described. However, before we look at this in more detail, there are two other factors which must be mentioned, both of which occurred in the late 1950s. The first of these was the introduction of rechargeable nickel-cadium cells, which rapidly gained popularity in spite of their high cost, as equipment was transistorised. The second was a move towards the use of superhet receivers This occurred rather quietly in the course of the miniaturisation that accompanied the elimination of relays but, due to cost considerations, it was some years before super-regenerative receivers totally disappeared from the multi-control scene

It was against this background that proportional radio control began to appear on more than just an experimental basis.

Probably one of the first successful attempts at proportional control was the two tone pulse width system, developed by Walt Good in the U.S.A. around 1955. This system did, of course, employ valves and relays and so was bulky, but was later to develop into the "galloping ghost" method of control. Walt's system transmitted two tones, which were continually switched from one to another. The rate at which they were switched defined one command, and the relative time that one tone was present compared to the other, defined the second control function. The receiver separated the two channels, and operated relays which were used to drive geared down electric motors with spring assisted centring. Later developments were to use a single electric motor to separate the mark space and the rate variations. This, then, only differed from the present day galloping ghost in that instead of switching one tone on and off, two tones were switched alternately.

The development of these rather elementary systems followed along similar lines to that of reed equipment. Relays were eliminated by the use of transistors, and performance was improved by raising the engineering standard of the mechanical parts. It is a little sad that a lot of experimental work produced unsuccessful results due to the somewhat crude mechanics often employed, and only in the U.S.A. was the market sufficiently large to attract those with the correct resources. In recent years, the benefit of experience gained from feedback proportional, has allowed the galloping ghost system to progress to a high standard and produce excellent results within its own limitations.

Undoubtedly there were other experimenters, but the first feedback proportional system to become widely known was the Space Control outfit, developed in 1959-1960 by Toomin and Richie in the U.S.A. This was a four function, fully proportional, system with closed loop servos, and offered a control effect similar to the simplest military systems. Many firms on both sides of the Atlantic started development programmes, and some emerged with a greater degree of success than others.

The earliest feedback proportional sys-

tems employed a transmission system similar to the "galloping ghost" technique, except that two tones were used as sub-carriers to allow four control functions to be performed. The servo amplifier circuits replaced the mechnical decoding of the control signals, and the spring centring was superseded by a potentiometer, whose wiper contact was coupled to the servo output. By applying a voltage across the potentiometer, the voltage output at the wiper accurately defines the position of the servo output, in a form that electronic circuits can accept. Hence the control accuracy was greatly increased over galloping ghost, to an extent that operating performance equalled the precision of reed systems, with the added advantage of greater smoothness.

Systems of this type were loosely termed analogue and suffered from a major problem. This was that, as range increased, the received signal suffered from distortion, resulting in a drifting of neutrals. There was also another problem, associated with centring accuracy, inherent in the type of servo amplifier employed. It was these problems that prevented proportional having an immediately superior performance over reed systems. However, proof that it was at least equal came in 1963, when Dr. Ralph Brooke of the U.S.A., tied for first place in the World Aerobatic Championships, using a commercial proportional system.

By the following year, at least in the U.S.A., proportional was fully accepted and nine of the top ten in the U.S.A. National Aerobatic Championships, were using this type of equipment. This rapid increase in popularity, was accelerated by the introduction of digital systems, which resulted in the elimination of the servo neutral variations inherent in the early analogue systems.

This was the thin edge of the wedge for reed systems. Once proportional reached a stage of affording a greater degree of control than reeds, aerobatic competition rules were changed to make manoeuvres more difficult, and a similar situation was later to occur in boat competitions. With this incentive, the demand for proportional increased, accelerating the rate of development by manufacturers.

With few exceptions, development was concentrated on digital systems since, initially, these appeared to offer greater potential. One subject that rapidly became evident as requiring attention was the superheterodyne receiver. The superhets that had been employed with reeds were found not to be satisfactory with proportional. This was because the reed bank effectively provided a further degree of selectivity, eliminating a high proportion of unwanted signals that had not been rejected by the receiver itself. It is fair to say that development along this line is still continuing, both with a view to increasing reliability, and to allow operation of an increased number of models at the same time.

Coupled with the increase in the radio link reliability, was the omission of failsafe circuits. On this type of equipment a failsafe facility can only guard against loss of radio contact. Since it must obviously be interconnected with the servo circuits, the risk of a malfunction in the failsafe increases the risk of an overall failure. In any case, with a highly manoeuvrable aircraft a large amount of damage is likely to result even in a failsafe condition. Thus, by omitting the failsafe circuits, nothing of value was lost, and a reduction in both volume and cost resulted.

With a digital system, the receiver is followed by a decoder, whose purpose is to separate the information for each control function from the incoming signal. This is described in detail in a later chapter, but suffice it to say that the type of circuit required is similar to that found in many applications of computer electronics. The early systems used decoders constructed from individual transistors, but it was not long before some manufacturers started to take advantage of components used in computers. A decoder constructed from computer logic is very compact, has increased reliability and, from the production viewpoint, is cheap to assemble and test. The cost of the components themselves is, however, higher and so not every manufacturer has adopted their use. As prices fall, and they are sure to with increased computer production, we can expect to see a greater trend towards the

use of these integrated circuit logic elements.

Meanwhile, there is another component which has become widely used for decoders This is the silicon controlled switch (S.C.S.). Decoders constructed using these and the associated other discrete components, are far more compact than the early transistor types and have resulted in a reduction in overall volume. Following on from here, an S.C.S. can be simulated circuit-wise by two transistors of very low price, and several manufacturers are using this form of decoder. This may, at first, seem a retrograde step, but the decoder operates in a different manner to the earlier transistor types and also employs fewer other components. This, coupled with the smaller physical size of modern transistors, allows a reliable, compact decoder, which, from a commercial consideration, has some economic advantages. It may well be a considerable period of time before any particular type of decoder gains a lead on the others.

The object of the continuing development would seem to be to gain increased reliability with a reduction in bulk, without sacrificing performance. It is fair to say that the demands of aircraft modellers are promoting this. The requirements for boat operation are no less stringent performance wise, but generally the overall size is of less importance.

In the majority of aircraft four servos are employed, one on each of the primary controls. Consequently, if the bulk of a servo can be reduced, it offers a significant improvement in the total complement of equipment in the model. It is not surprising, therefore, to find manufacturers concentrating on this area.

Since servos almost universally employ plastic mouldings, as the mechanics become smaller control of variations due to moulding techniques assume greater significance if an inferior performance is not to result. The cost of this type of work precludes all but the larger manufacturers, so we find that there are only a comparatively small number of different types of servo for all the equipment producers. As a result, we can expect that future progress in this respect will continue at a much slower rate, and will be dictated by a relatively small number of manufacturers.

Coupled with the miniaturisation of servos, has been the development of servo amplifiers. With smaller servos, the space available for the amplifier is reduced, forcing some circuit changes to be made. The first step was to replace some trigger circuits with high gain amplifier stages. With careful selection of transistor types the performance was not impaired, and a significant saving in the total number of components was achieved.

With the aim again of reducing the total number of components in the amplifier, the incorporation of integrated circuits was attempted, with a reasonable degree of success. Unfortunately, the servo amplifier does not follow very closely to standard computer circuits, so the integrated circuit is not employed quite in the manner on which it was designed to operate. This introduces problems in that the circuit performance varies with integrated circuits of different manufacture, even though they have identical specifications for their intended application. Problems of this type are obviously best avoided from a manufacturing viewpoint, but are by no means insurmountable

Several manufacturers have now overcome many of the problems associated with servo amplifiers, by having integrated circuits specially made to their requirements. By so doing it is possible to incorporate most of the extra components required for the circuit, with the result that the entire amplifier is one integrated circuit with only the timing components and possibly the output drive transistors being external from the encapsulation. The increase in reliability and the saving in manufacturing time has obvious attractions, but the initial investment is a considerable expense.

That is the picture to date as far as digital systems are concerned and, as in all commercial enterprises, some manufacturers can be expected to gain a lead over their competitors, as they achieve a lower price without sacrificing reliability. One cannot, however, ignore analogue systems. It was said earlier that they lost favour mainly due to servo performance, but the

development work on digital has pointed to some of the solutions to the problems.

Improvements in the performance of superhets went some way towards reducing servo neutral shift with signal strength, but the problem was really eliminated by changing to different forms of transmission. It is the mark-space variation that is most affected so, by using frequency variations only, the difficulty is overcome. We shall look at this in more detail later.

Difficulties inherent in analogue servo amplifiers have really been solved by development in transistors themselves, i.e. better transistors have become available at lower prices. This, coupled with a vast improvement in the quality of electric motors available, has resulted in analogue servos having a comparable performance to digital types. It is doubtful if analogue servos will ever become as miniaturised as digital types, since performance is far more dependent upon the servo mechanics, and the mechanical tolerance that will need to be maintained, will have a prohibitive cost from a commercial viewpoint.

Which way development is likely to proceed in the future is something that will become more obvious after discussion of the technical aspects in the coming pages. However, one thing is certain, modellers will continue to demand smaller equipment, and where there is demand, supply follows. We have already seen a reduction in size of servo motors resulting in a reduced all up weight for a multi-function proportional system, and circuit construction is making more extensive use of integrated circuit elements to reduce manufacturing costs and increase reliability. Several manufacturers are now moving in this direction for the superhet receiver itself.

However, short of a major breakthrough of equal significance to the introduction of transistors, it is hard to envisage any major changes occurring except to introduce more sophisticated transmission techniques. One area in which work is now being carried out is in the use of FM (frequency modulation) for the basic modulation system, and this should result in both increased reliability and more efficient use of the frequency bands allocated for model radio control.

CHAPTER 2

WHAT IS PROPORTIONAL?

THE object of fitting any model with a remote control system, is to allow the operator to make it perform manoeuvres under his command. The degree of control required is dependent upon just how precisely the operator requires the various functions to be performed, consistent with the model's own capabilities. Quite obviously one would not fit a single channel bang-bang system to a racing model aircraft flying at 100mph. Equally, a slow moving cabin cruiser model boat does not require an infinitely variable proportional system on anything other than, perhaps, steering. Thus, the type of control system is closely allied to the nature of the model it is to operate. We will consider aircraft as an example, but the following discussion also applies closely to both boats and land vehicles.

Consider that we have a model flying along a straight level path, and it is required to make it turn smoothly through 90° without losing any height. We have a control device which operates on the rudder only, and can give us either neutral or full throw in either direction, i.e. a simple bang-bang system. If we were to attempt to perform our turn by applying full rudder and holding this command for a period of time we would, undoubtably, meet with little success. This would be for a number of reasons, including the aerodynamic design of the model, and outside factors such as wind conditions.

It is immediately obvious that what has occurred is only a poor approximation of our required manoeuvre. What should have been done was that, from the moment the model started to obey our instructions, we should have applied corrections to compensate for all the deviations from the required flight path as soon as they became apparent. Let us assume that, by applying full rudder, the model enters the turn, but far tighter than we desire. This we can only correct, with our simple system, by applying full rudder in the opposite sense. A definite period of time will have elapsed before we initially observe that some correction is necessary and a further delay will exist before our corrective action takes effect. But, having made the corrections, we now have the case where the model will, approximately, follow our desired course. The flight path can be improved further by increasing the response times of the system, but very little can be done to improve the time period that elapses before the operator visually notices a deviation, although a certain degree of anticipation comes with experience. An even greater improvement is possible if one is using a reed control system since, with this, it is possible to "pulse" the switches, so producing an intermediate degree of control.

Returning to our example. The aerodynamic properties of aircraft are such that, if rudder is applied, although a turn results, the nose also drops. Hence we need to be able to apply elevator control at the same time as rudder, in order to maintain constant height. This, again, presents the same control problems as for the rudder, even in a simple manoeuvre such as a turn. With a bang-bang system smooth control is virtually impossible and even a skilled operator using reeds, manually pulsed, will only produce

relatively smooth results with a very slow, easy to fly, model. One does not need much imagination to appreciate the great difficulties that will be encountered as the model's performance is increased.

If, therefore, model performance is to be increased, a greater degree of control, with greater precision, is required. A suitable system must act quickly, since the visual feedback of errors cannot be speeded up significantly, it must also be capable of simultaneous operations, so that there is no time delay in applying corrections from a subsidiary control (i.e. the control that is not primarily producing the manoeuvre) to the main function.

The object of proportional control is to offer the greater degree of precision required and, at the same time, to offer simultaneous operation of control channels. With this greater degree of precision the aircraft performance can be improved and, since it is now possible to demand fine control movements, the aircraft can be so designed to respond gently to minute changes in its control surface positions, and yet be sufficiently responsive to perform aerobatics when so commanded.

Galloping ghost, the earliest form of proportional, is primarily intended for relatively slow flying and stable types of aircraft. With this type of system the whole control surface, usually rudder, oscillates between its two extreme positions and control is obtained by effectively varying the time period that the rudder is on one side of the neutral position. If the oscillations are sufficiently fast, the flight path is smooth, since the inertia of the model averages out the mean position. A form of feedback is introduced by using a light spring, so that there is no tendency for the control actuator to drift off in the direction of the command.

The system may be further extended to give elevator control and a bang-bang auxillary command for engine throttling.

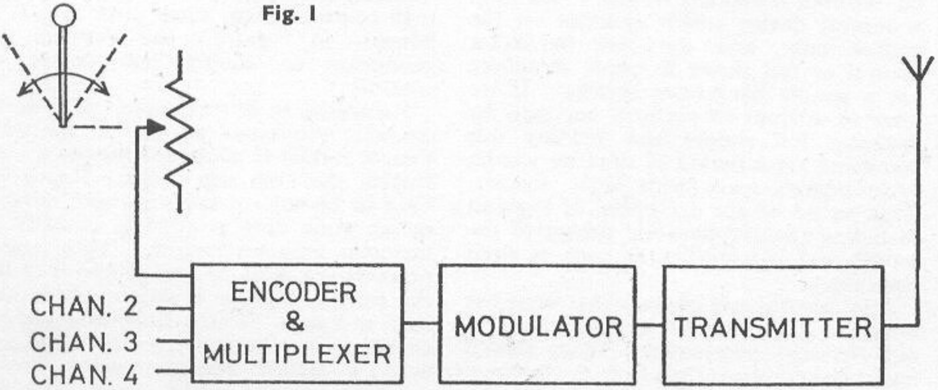
The limitations of galloping ghost are in its requirement for the model to average out the command signals and, hence, its dependence upon the model's characteristics. Feedback proportional is designed to eliminate this by using electronic circuits to average out the signals, and the servo systems' characteristics are virtually independent of the control surface and, hence, also of the model's flight characteristics.

A fully proportional system can, therefore, be defined as one in which several commands may be operated totally independent of one another, and in which the output devices assume positions defined entirely by the input devices, with negligible response time.

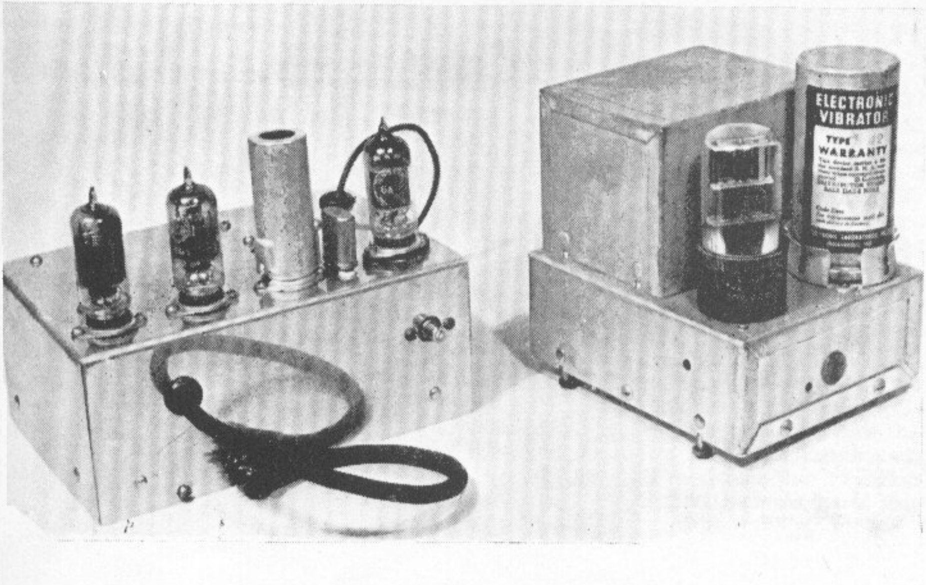
Basic systems

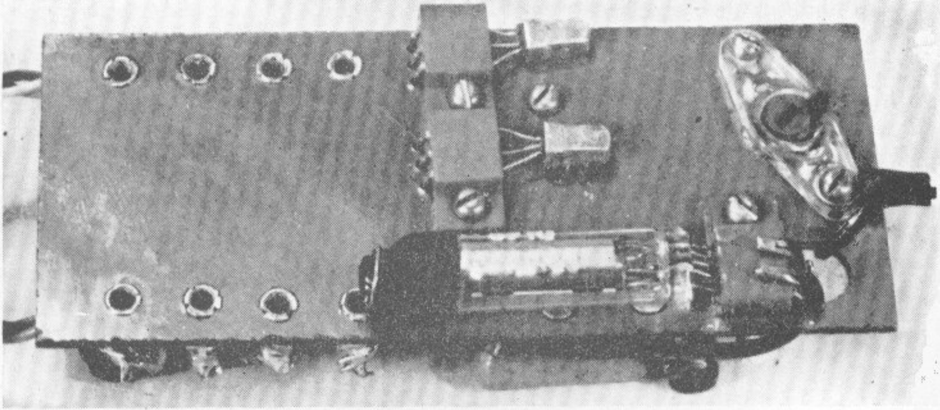
The basic functioning of any proportional system, whether analogue or digital is shown in Figs. 1 and 2.

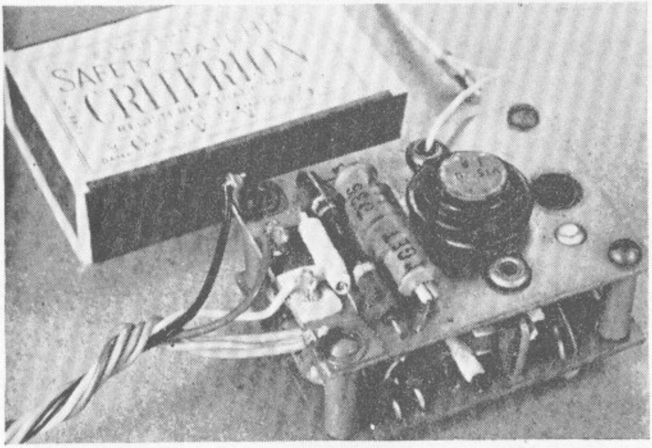
Fig. 1



Early proportional—" Galloping Ghost "—system with hand-held transmitter. Here George Honnest-Redlich carries a 6 volt scooter accumulator, slung from a shoulder strap, to power the transmitter via a vibrator, such as is shown below. These converted 6v. to 120v. for transmitter H.T.

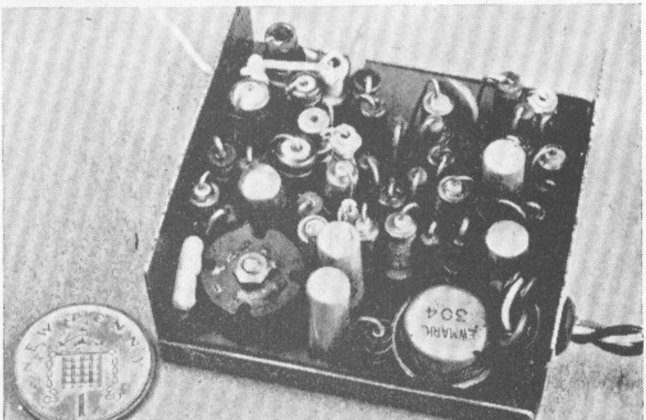






Above: an early "hybrid" receiver using a valve detector stage and two transistors for amplification

Left: one of the first "do-it-yourself" relayless, all transistor carrier receivers—the "Micro-dyne One"



Right: a relayless tuned filter all transistor receiver by British Ace

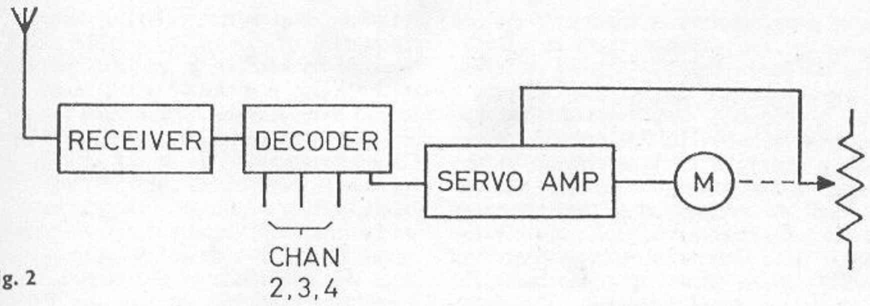


Fig. 2

Let us consider the transmitter, shown in Fig. 1. In general, an input command is defined by the position of a control stick, although other devices are sometimes employed for boats and land vehicles. The mechanical movement is converted into a variable voltage by means of a potentiometer coupled to the control stick, and this voltage is used to control an encoder. The purpose of the encoder is to convert the voltage to a suitable form for controlling the modulation of the transmission. In the majority of equipment the encoder also serves to multiplex the various channels in a multichannel system. The nature of the encoder-multiplexer is closely related to the type of modulation employed, and will be discussed in more detail later when the various types of transmission are described. Suffice it to say at this stage that the combined signal from all channels forms the input to the modulator section.

The modulator superimposes the channel information upon the radio frequency signal, and is designed in such a way as to prevent the transmission interfering with other signals on nearby frequencies. Thus we have converted several mechanically defined positions into a single signal that can be radiated from a transmitter.

Fig. 2 shows a receiving system, and it will be seen that the individual commands are separated in the reverse manner, to that by which they were combined. The receiver demodulates the incoming signal to a form similar to that which existed at the imput to the modulator in the transmitter. The signal at this point contains all the information for the different commands, and the radio frequency has been removed. The decoder accepts this signal and serves to separate the information for each channel onto an output for each servo amplifier. Finally the servo and its amplifier convert the signal back into mechanical movement, a potentiometer coupled to the output shaft providing information to the amplifier to define the position of the output shaft at any time. By comparing the incoming signal with the feedback signal, the servo amplifier determines the difference between the demanded position and the present position and drives the motor accordingly, until there is no difference. In this condition, the demanded position and the output arm position are identical and the servo has reached the desired deflection.

Modulation

The principles employed for transmitting information to a radio controlled model are exactly the same as in other branches of communications. We have a single communication channel, a radio frequency signal of around 27 megacycles, and this has to be used to convey the information for a number of control functions. This is achieved by modulating our carrier signal in some manner with the appropriate information. There are three types of modulation possible, amplitude, frequency and pulse.

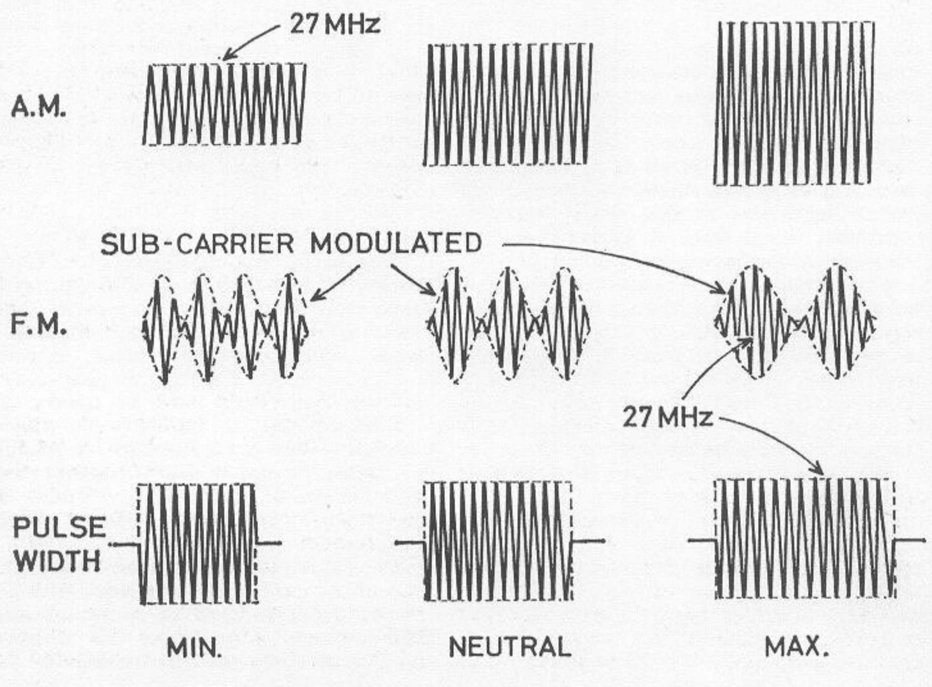
Amplitude modulation varies the amplitude of the carrier in accordance with the information that is to be communicated. The receiver has to detect this variation and this is where the disadvantage of this

type of system becomes apparent. As the distance of the receiver from the transmitter is increased, the signal strength decreases. Hence the receiver detects a signal of reduced amplitude and has to determine whether this is caused by the distance changing or is a change in the channel information. Circuits can be designed to average out the changing channel information, and if the mean value is used as a reference, the gain in the receiver can be varied to compensate for the changing signal strength. The complication of this compensation virtually eliminates the use of this otherwise simple system for model control purposes.

Frequency modulation systems vary the frequency of a signal in accordance with the control information. In the receiver, provided the minimum acceptable signal level is exceeded, the transmitted frequency can be recovered without any inaccuracy being introduced by the communication link. At radio control frequencies it is not practical to vary the frequency of the R.F. signal. This is because to achieve a realistic percentage change, the frequency would need to be varied over a wider range than the whole of the permitted radio control band (26.96 to 27.28 MHz in the U.K.). Also at these frequencies the circuit design becomes more difficult. To overcome this problem a tone, usually in the audio range, is used as a sub-carrier. This tone modulates the amplitude of the carrier, but the amplitude does not convey any information. If the tone frequency is varied this can be used to carry the control information. It is particularly important to maintain a high degree of stability in the tone generators since if the control information is specified by a 10% change in frequency, an inaccuracy of 0.5% in the frequency results in a 5% positioning error at the servo.

Pulse modulation is the most widely used type of transmission for proportional

Fig. 3



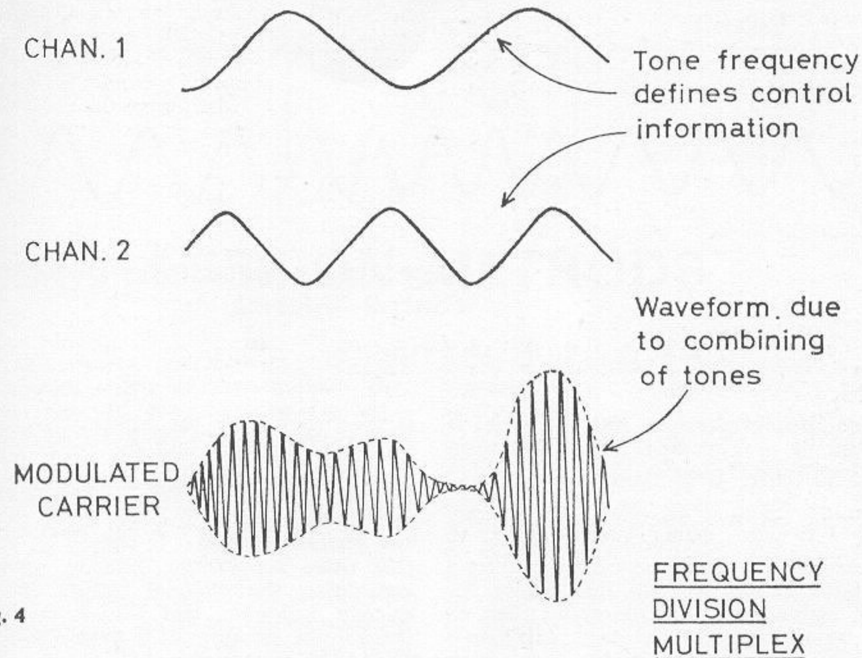


Fig. 4

control. Various types of pulse transmission are possible but we shall only concern ourselves with those that are dependent upon pulse length. The galloping ghost system uses a variation in the ratio of the duration that the pulse is on to the time it is off. This is termed a mark-space variation. A similar transmission can be employed for feedback proportional systems. A variation of this is employed in the transmission system commonly employed in digital proportional. In this application the required servo position is defined by the duration of the pulse and is not related to the off period. A narrow pulse indicates one extreme servo position, and a wide pulse the other extreme. This is known as pulse width modulation.

Fig. 3 shows the forms of the transmission signals at both extreme servo positions and neutral, for each of the three types of modulation described. It will be noted that both amplitude and frequency modulation continuously specify the required servo position. With pulse width modulation, the servo position is only specified at discrete instants of time, i.e. when the pulse ends. It is therefore necessary to repeat the pulse at frequent intervals so that there are no apparent gaps in the time taken for the servo to respond. In digital proportional systems the pulse is repeated at intervals in the order of 20 milli-seconds, which is obviously far in excess of the rate at which the servo can respond. Consequently there is no apparent discontinuity in the servo response.

Command Multiplexing

Our definition of feedback proportional has specified that several functions must be capable of being performed simultaneously. We have also seen that our transmitted signal has to convey precise information regarding servo position. Clearly it is not practical to use separate transmitter frequencies for each of the required control functions, so we must code our signals in such a way as to permit all the information to be transmitted on a

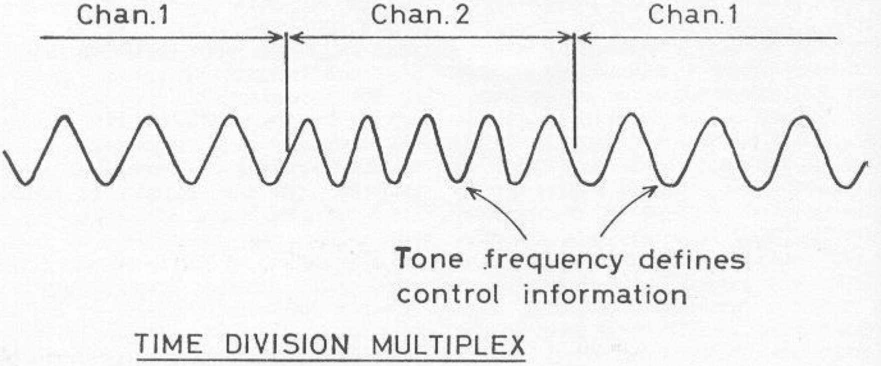


Fig. 5

single communication link. This is known as multiplexing the signals from the command channels.

There are two methods of achieving this. If our command information is defined by a tone frequency, we can use a different range of tone frequencies for each command. The tones can then all be transmitted continuously. This is termed frequency multiplex since each command is specified by a particular frequency range. Alternatively time multiplex may be employed. For this, a given time period is sub-divided into smaller intervals. Each of these intervals is used to convey the information for a particular channel, and the entire frame of channel information is repeated at frequent intervals. Figs. 4 and 5 show two possible configurations for a simple two channel system. It should be noted that there are several different ways of combining the channel signals in frequency multiplex, and Fig. 4 has taken the specific example of a simple addition of the two waveforms. It is also apparent that for pulse width modulation, a time division system is the simplest.

In the next chapter we shall consider the various types of transmission methods in greater detail.

CHAPTER 3

TRANSMISSION METHODS

IN this chapter we shall consider the different types of transmission that can be employed for multi channel systems, and look at the merits of each type. Before we do this it is perhaps necessary to give the reader an idea of what a radio frequency signal is and to explain some of the terminology.

The Carrier Signal

The radio frequency signal radiated by our transmitter aerial carries information to the receiver. This is achieved by a transfer of energy. To explain how this occurs we will consider an analogy with sound. If we have a loudspeaker the cone of which is vibrating, the air adjacent to it is compressed as the cone pushes forward, and rarefied as the cone pulls back. If there is a local increase in the air pressure, air attempts to flow to regions of lower pressure. Since a finite amount of time is taken for the air to physically move, before the pressures have fully equalised the cone has retracted and produced a reduction in pressure. Hence the air attempts to flow back to its original position. Thus as the cone vibrates the surrounding air is caused to vibrate and this in turn causes the air further from the loudspeaker also to vibrate. It should be noted that whilst the air is in motion, it does so in alternate directions and consequently its mean position is unaltered. Thus there is not a general movement of air away from the loudspeaker. What does occur however is a transfer of sound energy from one particle of air to the next. Since a certain amount of energy is expended in setting the air in vibration and also to maintain the motion, the further the observer is away from the loudspeaker, the smaller the amount of energy that reaches him.

The transmission of radio waves is very similar but instead of energy being transferred by the vibrating air particles, it is oscillations of an electromagnetic field that propogate the signal. Light is a form of electromagnetic radiation, the oscillations occurring in a frequency range that is visible, different colours being caused by different frequencies of vibration.

This gives a clue to where electromagnetic waves differ from sound waves. Sound requires the presence of particles in order that it may be propogated, and the speed of propogation is dependent upon the nature of the particles. Thus the speed of sound in air differs from that in water. Electromagnetic waves are independent of any form of medium. Hence their speed in air is the same as in a vacuum, and it is the speed of light. Providing the reader appreciates that there is a field of influence around a radiating source, and that it is energy that is propogated, there is no need to consider the complex theory of electromagnetic fields.

One method of inducing a vibration into an electromagnetic field is from an electrical circuit across which the voltage is changing. In a transmitter we cause a circuit to oscillate at the required frequency i.e. the voltage across the circuit varies in a cyclic manner. By connecting a suitably proportioned aerial to the circuit, the oscillations cause the voltage in the aerial circuit to oscillate at the same frequency. The aerial then acts in the same manner as the loudspeaker and transfers energy in regular bursts to the electromagnetic field.

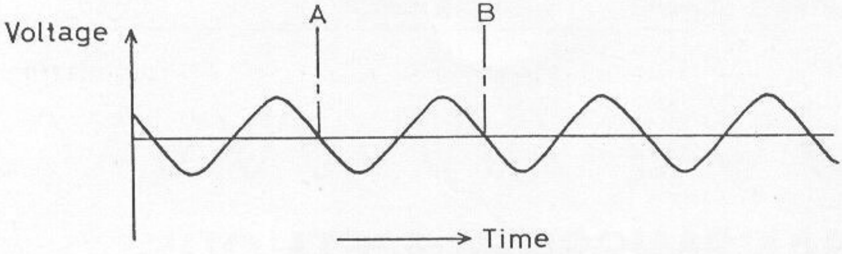


Fig. 6

The frequency at which these bursts occur is known as the radio frequency and the energy propagation is the carrier signal upon which our control information is impressed.

To detect a signal, a receiving aerial is placed into the electromagnetic field in such a way that it accepts some of the energy reaching it. The receiving aerial circuit therefore has an output in the form of a voltage which is varying in the respect to time. Fig. 6 illustrates this. The time that occurs between points A and B represents the period for one complete cycle of the oscillation. Periodic time is obviously related to the frequency and is simply the reciprocal, i.e. the periodic time is determined by dividing the frequency into one unit of time. Thus 27 MHz (27 million cycles per second) gives a periodic time of 0.037 millionth of a second (37 nanoseconds.)

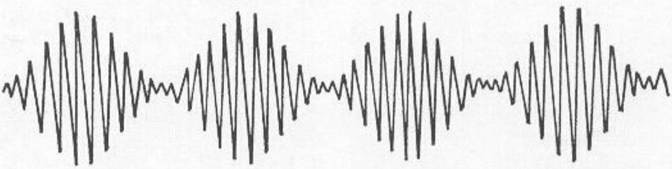


Fig. 7

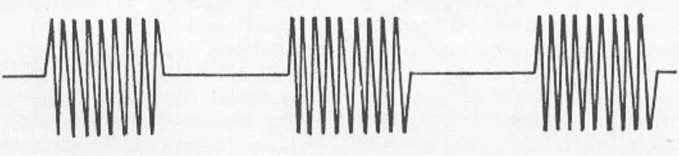
same manner as the voltage driving the transmitter aerial.

Waveforms

Since the electromagnetic field is not tangible we can only observe its presence by its effect upon other devices. Thus we measure field strength by means of the voltage that the field induces in a simple receiving circuit. The waveforms depicting signal information are therefore graphs showing the variations of voltage with

If we were to observe the voltage at the input to our transmitter aerial on an oscilloscope, we would see a trace as shown in Fig. 6. Similarly we could observe the signal at the receiver. The overall shape should be the same although the amplitude of the voltage would be considerably reduced. (There are difficulties in practice when observing signals at high frequency and these will be discussed in a later chapter. This applies particularly where the signal is very small.)

Fig. 8



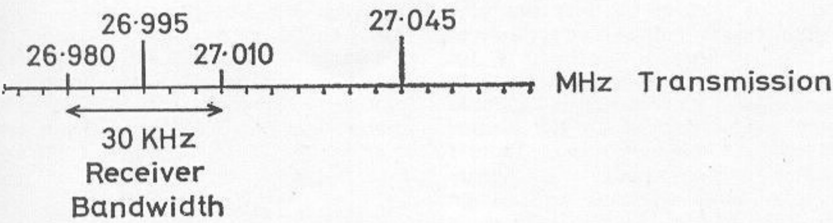


Fig. 9

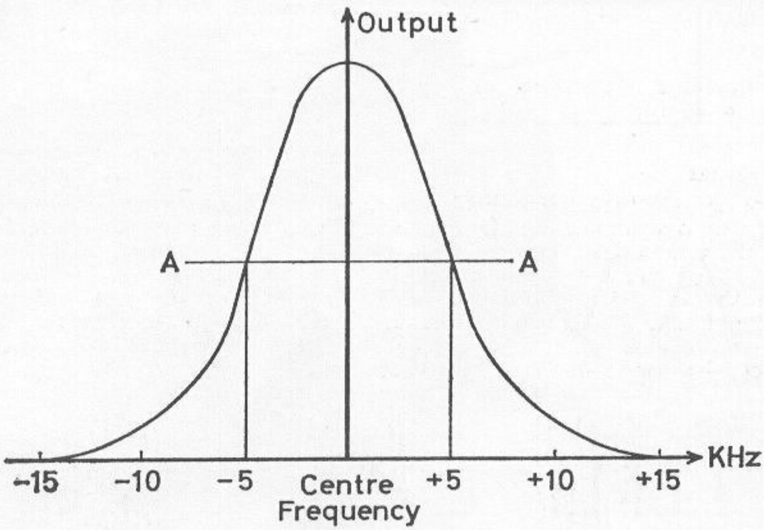
We have previously discussed the need to modulate our carrier signal in order to superimpose information upon it. For proportional radio control applications we use pulse modulation of the radio frequency carrier, and our channel information is defined by time variations of this modulating signal. A high frequency carrier wave modulated with a low frequency tone is shown in Fig. 7, and with a pulse waveform in Fig. 8. Obviously the waveforms for a multi-channel system become more complex than these two examples, but the principle is the same.

Bandwidth

Any device that can respond to variable frequency inputs does so only over a limited frequency range. The human ear can typically respond to frequencies within the range 18Hz to 16KHz, and a servo system can only follow the demand signal if it does not vary at more than a few cycles per second. The frequency range over which a device can respond is called its bandwidth and is specified for input signals having a sinusoidal waveform. Perhaps the most frequently encountered example of bandwidth in radio control is that of the receiver. If we have a receiver that is intended to respond to a transmission on say 26.995MHz (brown frequency) it must not be affected by a transmission on 27.045MHz (red frequency). Fig. 9 shows that the receiver must have a sufficiently narrow bandwidth for the unwanted transmission to be outside its pass band.

So far we have defined bandwidth as the frequency range over which a response is

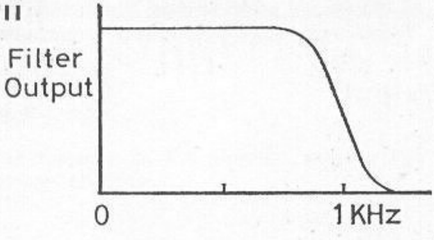
Fig. 10



produced. In practice the response of a system falls off gradually on either side of its nominal frequency as shown by Fig. 10. Because of this, it is more usual to state the bandwidth corresponding to the frequency range over which the output exceeds half its maximum value. Thus if we have a receiver with a frequency response such that its signal output voltage varies as in Fig. 10, and this operates a decoder which requires an input signal to exceed a threshold level at point A, the effective bandwidth of the system is reduced to 10KHz.

In the case of a receiver, the narrower its bandwidth, the better its selectivity. However, apart from stability problems, the system must have sufficient bandwidth to allow the required signal to be passed. Included in our definition of bandwidth is the fact that the input signal is sinusoidal. If we have a filter circuit which is designed to pass signals from 0 to 1KHz, i.e. a 1KHz bandwidth, a 1KHz sine wave will be passed without any change in its shape. If we were to attempt to pass a 1KHz square wave, the output would also be a sine wave. This is because there is insufficient bandwidth to pass the signal

Fig. 11



correctly. Fig. 11 illustrates this.

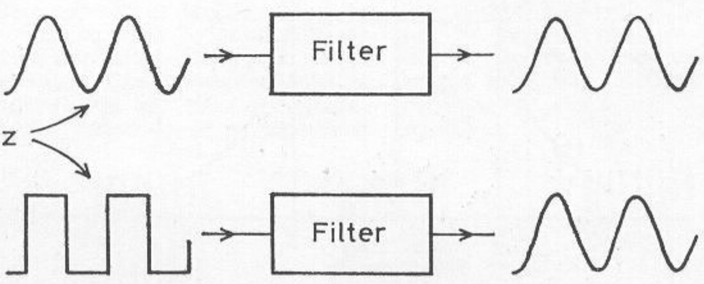
It can be shown mathematically that any waveform can be produced by adding together sine waves of the fundamental frequency and its harmonics in appropriate proportions. The series for a square wave of frequency f is shown in the following table.

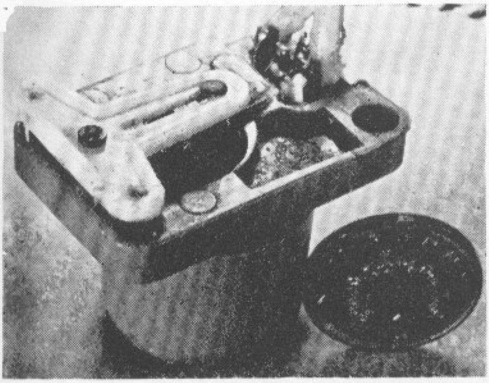
Sine wave amplitude	Sine wave frequency
1	f
1/3	3f
1/5	5f
1/7	7f
etc.	etc.

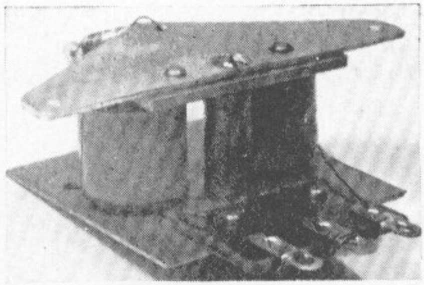
Fig. 12 shows the result of summing the first three frequencies and it can be seen that an approximation to a square wave is produced. A better approximation can only be achieved by taking into consideration a large number of harmonics. The waveform produced by only considering frequencies up to 5f, represents the type of response that would be produced if a square wave of frequency f were passed through a filter which rejected all frequencies greater than 5f. Obviously the filter network has insufficient bandwidth to pass the signal with any degree of accuracy. Thus when deciding upon the form of signal which will be used to transmit information, the bandwidth required in the receiving system must be taken into consideration.

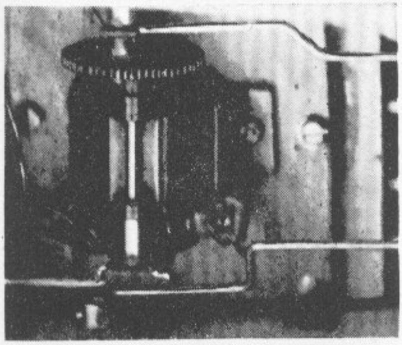
Frequency Multiplex

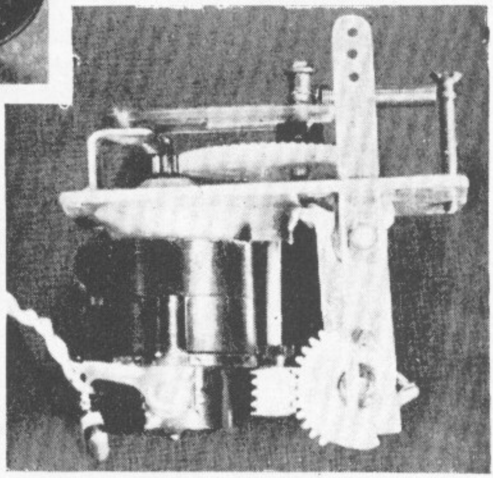
In a frequency multiplex system the control information is continuously applied to a separate sub-scarrier for each channel. The available bandwidth in the communi-

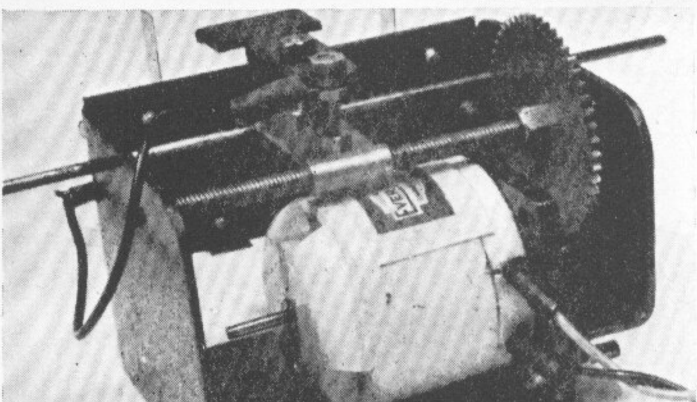






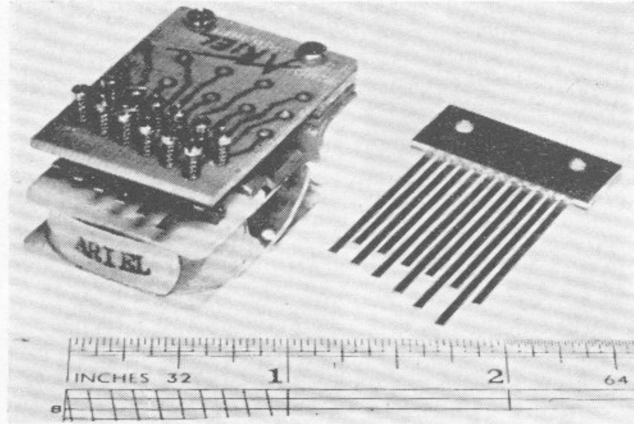






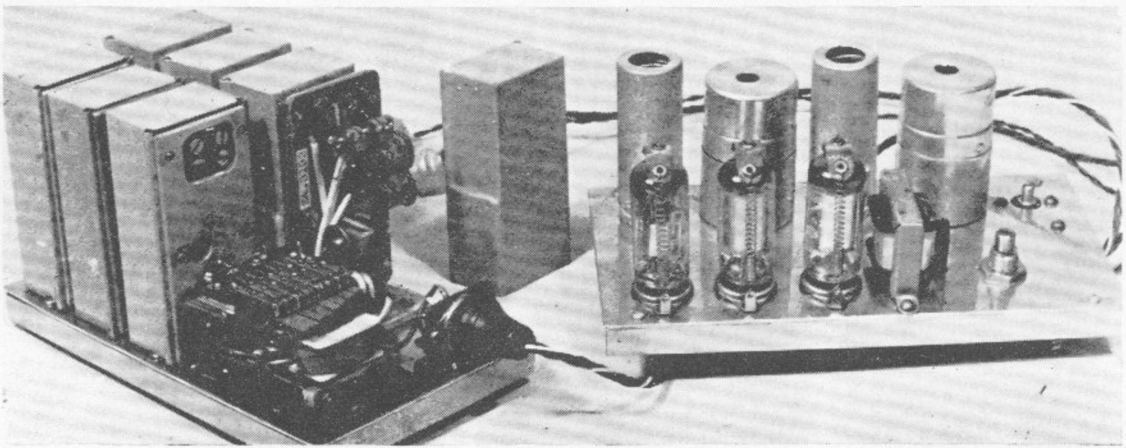
A selection of early actuators (servos). Top left the Mk. I Bellamatic, used on some multi channel outfts, had spring centring. Top right: a spring centred magnetic rudder actuator could be pulsed or operated by two relays. Centre: popular Mighty Midget motor was the basis for early galloping ghost actuators.

Left: an early servo for multi channel control using a lead screw and control switching tracks

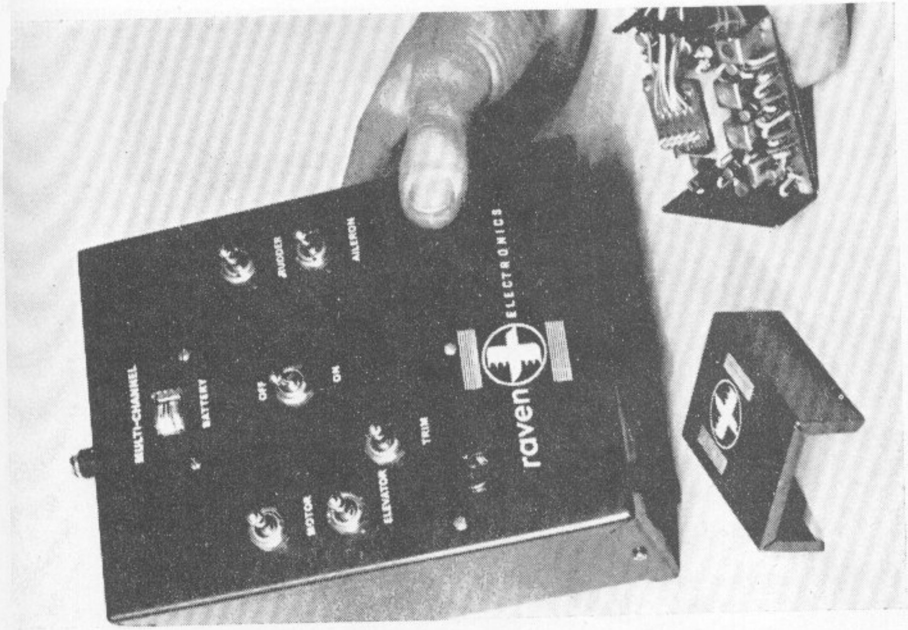


Reed bank function selector devices came in various forms. That illustrated on the right is interesting in that adjacent reeds are not tuned to adjacent tones, the idea being to prevent " air drive " pulling adjacent reeds and giving an unwanted control

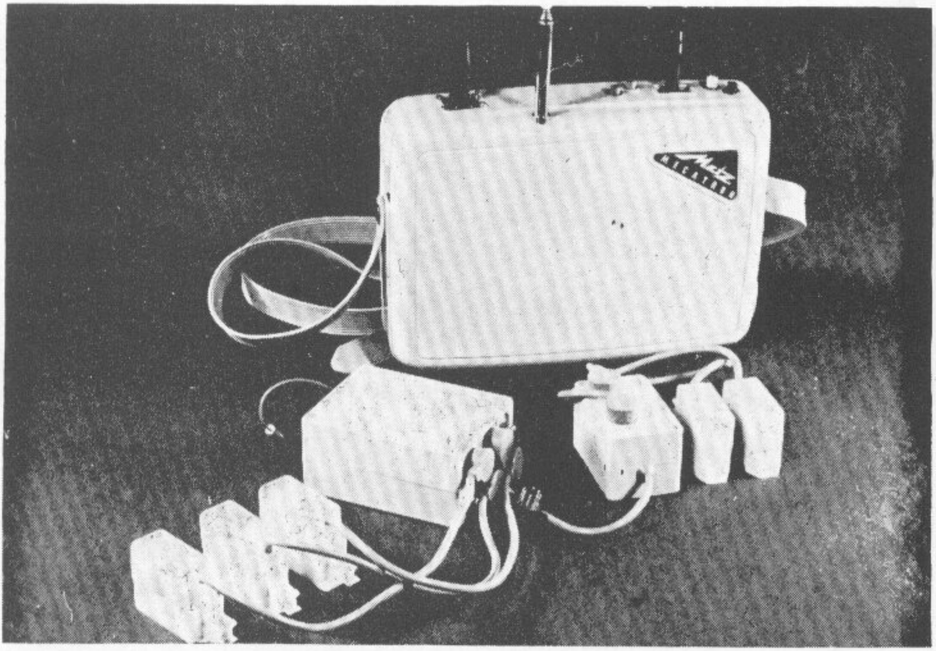
Below: How's this for size ? An early superhet with reeds and relays, probably intended for use in boats, has 6-channels (three functions)



Facing page: two more recent reed outfits—a valve/transistor hybrid super-regen by Orbit, and an all transistor superhet by a British manufacturer—Raven


MULTI-CHANNEL
BATTERY
OFF
ON
MOTOR
ELEVATOR
TRIM
RUDDER
AILERON
raven
ELECTRONICS


ORBIT
ORBIT



Tuned filters were used in the Metz outfits (above), which is interesting in that the servo had twin commutator motors to simplify the switching stages of this solid state outfit. Add-on units provided further filter outputs for two more servos, to convert the basic " three " to a " five." The Grundig outfit (below), also featured plug-together units, so that an outfit could be " built-up " in stages

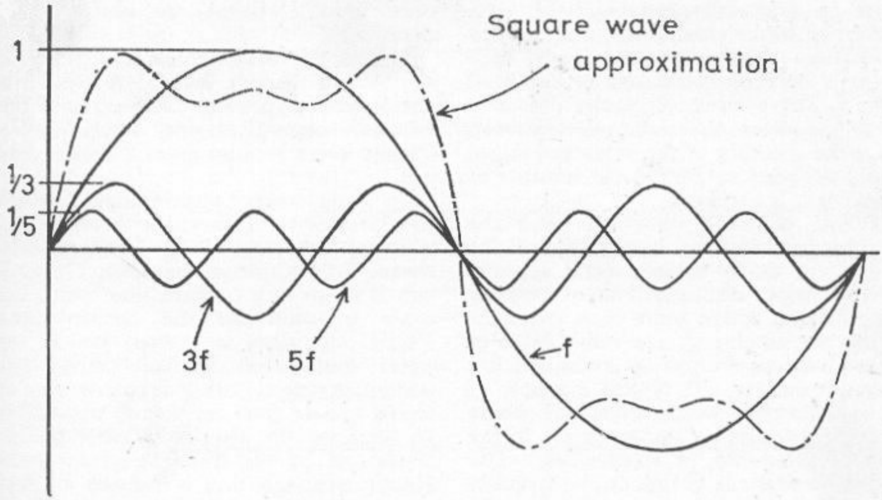


Fig. 12

cation link is divided between the sub-carrier frequencies. A typical radio control superheterodyne receiver with an I.F. (intermediate frequency) of 455KHz will have a useable bandwidth of about 6KHz. Hence for a four channel system, sub carriers of 2KHz, 4KHz and 5KHz could be employed. However when the frequencies are mixed together, as occurs in a transmitter modulator, various cross modulation frequencies are generated e.g. when a 2KHz signal is mixed with one of 3KHz, strong components are generated at the sum and difference frequencies (5KHz and 1KHz), and to a lesser degree at the other frequencies. In this example, an error would be caused in the sub-carrier which is operating at 5KHz. Similarly the difference frequency generated by the 5KHz and 3KHz sub-carriers will affect the 2KHz signal. Other interacting combinations should also be apparent.

Another aspect that must be considered is the effect of harmonics. Even if the sub-carriers were perfect sine waves when originally generated, some distortion will invariably have been introduced elsewhere in the communication link. Thus the 2KHz sub-carrier would contain frequency components at its harmonics. Consequently the second harmonic will cause an error in the 4KHz sub-carrier.

A more satisfactory choice of sub-carrier frequencies would be 1.6KHz, 2.5KHz and 5.6KHz. Clearly with the restrictions imposed by the system bandwidth, the choice of sub-carrier frequencies becomes increasingly difficult as the number of channels is increased. Allied to this is the frequency response of the filters used to separate the sub-carriers in the receiver. The bandwidth of each filter must be sufficiently large to allow the signal information contained in the sub-carrier to be passed, and also to accommodate any inherent drift in the frequencies from their nominal values. It must however be sufficiently narrow to ensure that any output caused by the adjacent sub-carrier frequencies is below the minimum input signal level of the following circuits.

With a frequency multiplex system, assuming a fixed receiver sensitivity, the transmitter power must be considered. If a single sub-carrier is employed to modulate the R.F. (radio frequency) signal, all of the transmitted power is available to that channel. A square law relationship exists between the power in each sub-carrier and the number of sub-carriers. Thus if two sub-carriers are employed, the

power in each sub-carrier is only one quarter of its original level, and if four sub-carriers are employed, each sub-carrier is only one sixteenth of the total power. The remainder of the power is lost to the cross modulation frequencies. Hence the accuracy of the recovered signal rapidly becomes suspect as the number of sub-carriers is increased.

Because of these difficulties and the exacting performance demanded of the circuits, for model radio control applications, frequency multiplex systems become unsatisfactory where more than two sub-carriers are employed. In other fields of communication, frequency multiplex has many advantages. A typical example of this is in satellite communications where the radio frequencies employed are in the order of thousands of megacycles. The bandwidth of the R.F. link can be typically a few megacycles, so allowing many sub-carriers to be employed. The difficulty of decreasing power in the sub-carrier is overcome by using a beamed aerial to concentrate all the power onto the receiving system, and the relative cost of increasing the transmitter power is small compared with the total cost of the system.

Time Multiplex

In order to increase the number of channels that can be accommodated within a comparatively narrow bandwidth, a different form of multiplexing can be employed. Instead of dividing the available bandwidth between the channels, it is shared on a time basis. Thus the entire bandwidth is made available to each channel for a limited period of time. For model control applications, provided the channels are sampled at a rate in excess of about ten samples per second, there is no discontinuity in the overall control response. Consequently with the growing trend towards superhet receivers of greater selectivity and hence narrower bandwidth, time multiplex is invariably employed in both analogue and digital systems.

Analogue and Digital

The words analogue and digital describe two types of radio control system, and when used in this sense their meanings are only loosely based on their precise definitions.

Digital precisely defines a variable that changes in discrete steps. A clock with hands gives an analogue indication of time whereas one displaying figures which change every minute gives a digital read-out.

In radio control terminology, analogue describes a system where the servos are supplied with varying voltage levels to define the required output arm position. Generally this is allied to a transmission which uses tones to modulate the carrier signal. Digital describes a system which uses pulse modulation of the carrier, and computer type counting circuits so that the signal sent to each servo is in pulse form. In this case the required output position is defined by the duration of the pulse. Hence although this is termed a digital system, the time duration is varied in an analogue manner to effect the control commands.

Digital Systems

The majority of radio control equipment commercially available falls into this category and has its origins in the U.S.A. where the system was first developed. In attempting to overcome the difficulties associated with analogue servos, an alternative approach was adopted that required the servo amplifier to be driven by pulses of variable width. The nature of the control information which then had to be transmitted was such that both transmitter and receiver circuits became greatly simplified. Without doubt this was a major factor contributing to the success of digital equipment.

In the transmitter each control stick produces a pulse the length of which is related to the stick deflection. Since the channel information is to be time multiplexed, the pulses are generated sequentially. It is a simple matter for the generation of channel 2 pulse to be initiated by the end of channel 1 pulse, and so on down the chain. The whole sequence of events is repeated usually at about 20ms. intervals under the control of a master "clock" which also serves to initiate the channel 1 pulse. Fig. 13 illustrates this sequence. Since the trailing edge of one pulse

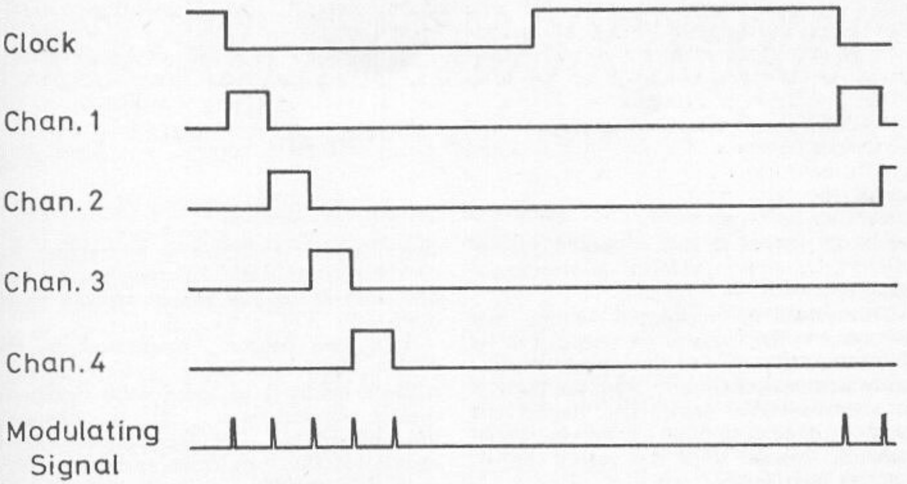


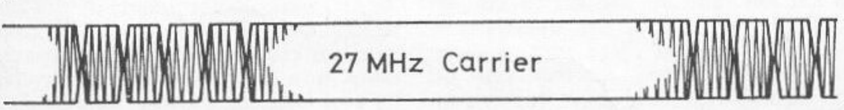
Fig. 13

occurs at the same time as the leading edge of the next, the multiplex channel information can be defined by a series of narrow pulses coinciding with these edges. This signal completely specifies all of the required channel information and may therefore be used to modulate the transmitted signal

In the receiver it is necessary to identify the beginning of an information frame in order that the separated command pulses are routed to the correct servo. The long gap occuring before the modulation pulse defining the beginning of channel 1 is used for this purpose.

The result of applying the modulation signal to an R.F. carrier is shown in Fig. 14. It is apparent that the rise and fall times of the modulation pulse will affect the accuracy of the command information. With good receiver design only one of the transitions will be important in this respect, i.e. the decoder will operate on only the leading edge or the trailing edge of the pulse. If it is assumed that the leading edge is to be used, the positive going transition must be fast for maximum control accuracy. It is therefore necessary to be able to switch the carrier signal off rapidly. Circuit design to achieve this can be difficult since the R.F. output stages of a transmitter are designed to resonate electrically at the carrier signal frequency. If a resonating circuit is suddenly switched, there is a tendency to produce spurious signals of other frequencies, resulting in "splatter." This effect is analogous to a swinging pendulum which is stopped by slowly moving in a hard object from one side, in the direction of the swing. The frequency is defined by the time required to travel from one extreme of its swing, out to the other extreme, and back again. The hard object upsets the conditions and the rebound caused by the pendulum bob striking it reverses the direction of the swing. The pendulum therefore reaches its original extreme position at a different time to that at which it would otherwise have done. The sequence continues with the time period changing every time energy is lost through the impacts with the hard object, until the pendulum

Fig. 14



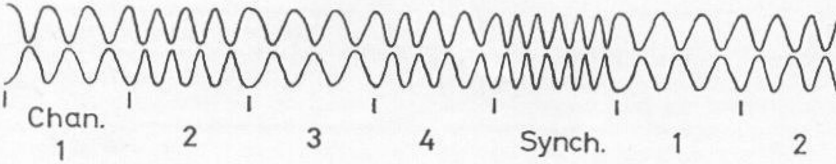


Fig. 15

eventually comes to rest. The frequency of the swing is therefore continuously changing.

The effect of unwanted frequencies appearing in the transmitter output can be to cause interference to equipment operating on adjacent channels. Also the correct receiver may have insufficient bandwidth in its tuning circuits to accept these transient signals, with the result that it operates incorrectly.

It should be apparent that there are conflicting requirements in the details of the communication link, with the result that the system has to be a compromise design. For maximum selectivity the receiver bandwidth has to be narrow, but for good accuracy the modulation has to be fast, requiring greater bandwidth. The need for modulation increases the frequency range occupied by the carrier signal, so necessitating a decrease in receiver selectivity. It is for these reasons that some early systems suffered from poor performance when operated alongside equipment of different manufacture.

With higher output from the transmitter the effect of splatter could become more pronounced. Hence a characteristic of the latest types of transmitter is the inclusion of circuits to specifically shape the modulation pulses. This will be discussed in more detail in a later chapter.

Analogue Systems

In analogue radio control systems the control information is conveyed by a tone. Information can be conveyed by both the frequency of the tone and its mark-space ratio i.e. the ratio of the on to the off portion of the cycle, if the waveform is square. Because of these two separate forms of intelligence, systems can be designed which use the signals in a variety of different manners. The two methods described are particularly interesting due to their completely different approaches, and also allow the major aspects to be discussed.

The first system, originated in the U.S.A., and now being used by a U.K. manufacturer, is in some ways similar to digital transmissions. All the channels are defined by tone frequencies, and a sample of the tone from each channel is sent sequentially. Synchronisation is achieved by sending a sample of a tone outside the normal command range, at the end of each sequence. Fig. 15 illustrates the modulation envelope of the carrier signal i.e. the R.F. carrier is omitted. This type of transmission allows the tone waveform to be other than square so reducing the difficulty of switching transients affecting R.F. channels. A transient problem could exist where the tone samples change from one command channel to the next, but with good circuit design techniques the problem is far less pronounced than when switching R.F. signals.

This transmission method offers many advantages due to it not requiring the use of square waves. If sinusoidal modulation is used, the receiver bandwidth does not need to be greater than twice the frequency of the highest tone. Potentially the design of systems having a bandwidth of only 2-3KHz is possible provided adequate stability of the critical components can be achieved.

From the transmission viewpoint this system is far less susceptible to malfunction caused by signal interference than a digital transmission method. With the latter, the channels are separated by counting the incoming pulses whereas the decoder in the analogue receiver generates its own stepping pulses. The extra pulses that might be introduced into the signal

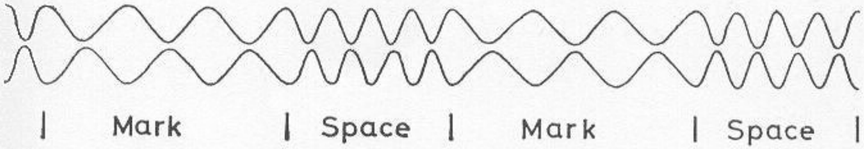


Fig. 16

by interference will only result in a slight drift of the servo from the correct position, and control will be reasonably maintained compared with the digital system.

One feature that both digital and this type of analogue system have in common is that each individual control channel is handled in the same manner. This results in the use of repetitive types of circuit and so minimises any differences in the performance of individual channels.

The other type of analogue system to be considered uses tones to convey information for two channels. As in the preceding system the frequency of the tones defines the required control position but the manner in which they are multiplexed differs. The transmitted signal switches between the tones so that they are alternately sampled. The modulation envelope is shown in Fig. 16. By varying the sampling rate and its mark-space ratio two further command channels can be included. In order that the receiver can separate the commands the signal has to convey some form of synchronisation information. This is achieved by using two widely different tone frequencies which can be separated by filter networks in the receiver. This requires the communication channel to have a greater bandwidth than for the other transmission described, but the system has the advantage that the information for each command channel is present for a greater proportion of the time. Theoretically this gives a greater degree of accuracy in the control information but at the expense of extra complication in circuit design. Probably the greatest disadvantage of this system is its lack of any expansion facility for increasing the number of channels for extra control functions. The circuits are relatively complex and do not readily lend themselves to miniaturisation by the incorporation of integrated circuits. Offsetting these disadvantages is the inherent advantage of an analogue system being able to maintain a reasonable degree of control in the presence of interference.

CHAPTER 4

DIGITAL VERSUS ANALOGUE

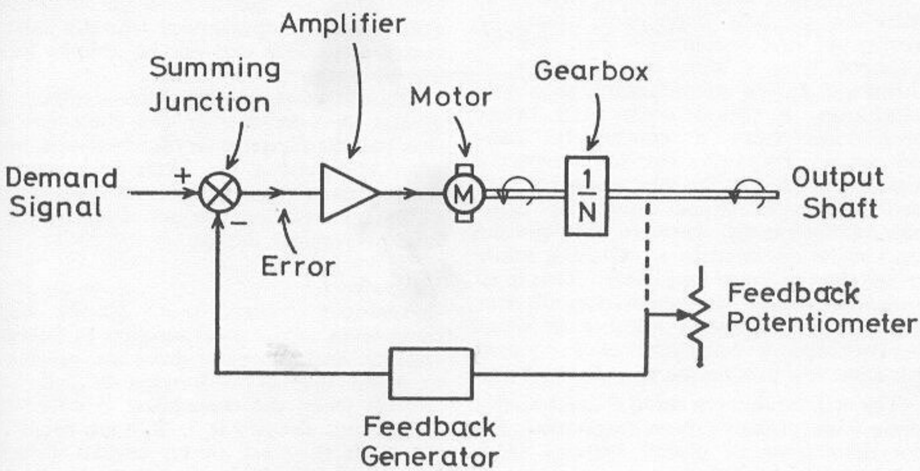
Servo Operation

The essential difference between analogue and digital radio control systems is the nature of the signal which defines the control information to the servo. Analogue receivers provide an output in the form of a voltage, the level of which varies in accordance with the demanded servo position. The input to a digital servo is in the form of a pulse and only the time duration conveys the necessary information. The voltage level of the pulse is completely irrelevant provided a threshold level is exceeded.

The functioning of a positional feedback servo is shown in Fig. 17, and applies to both analogue and digital types. For an analogue servo the feedback generator is a simple low gain amplifier providing a voltage output linearly related to its input from the potentiometer. The summing junction subtracts the voltage representing the present output arm position, from the demanded position voltage and provides an error output. The error may be either in a positive or negative sense depending upon which direction the servomotor has to drive to reduce the error.

For a small error signal, the output from the amplifier to the motor will be small resulting in low power output from the servo. Hence the accuracy with which a servo can take up a position relies to a great extent on having low friction in the gear train, and is dependent upon the load on the output arm. Since a certain voltage has to be applied across the servo

Fig. 17



motor to overcome friction before any motion begins to occur, a discrepancy, which cannot be corrected, can exist between the demanded and output positions. If the amplifier gain is increased to overcome this difficulty, the servo will tend to overshoot the required position resulting in oscillations of the output arm.

In a digital servo, the feedback potentiometer controls the duration of a pulse, which is generated by the leading edge of the incoming demand pulse. The summing junction compares the length of these two pulses and gives either a negative or positive output pulse whose length indicates the error. These error pulses are stretched so that except for extremely small errors, a continuous output is produced in the correct sense. This causes the amplifier to provide its full output voltage in the appropriate direction to the motor. For small errors the pulse stretching circuit provides an output in the form of a longer pulse, the length of which is related to the size of the error. This results in the motor being supplied with full voltage pulses of variable duration. The motor is thus pulsed in a direction that will tend to correct the error. The effect of this pulsing is to overcome most of the friction in the system and also the effects of stickiness in the gear train are minimised.

The performance of a digital servo is therefore far less dependent upon the quality of the mechanics than for an analogue type. With the problems of shrinkage during manufacture of plastic mouldings, a digital servo is a better proposition from a commercial consideration. From the electronic aspect a digital amplifier is far less susceptible to performance variations caused by component tolerances, because its circuits tend to be operated in a switching mode rather than as linear amplifiers. This is all important where miniaturisation dictates the use of a minimal number of components and a minimum of individual attention to a production run of amplifiers.

These differences in servo characteristics alone have probably been responsible for the dominance of digital systems from commercial manufacturers.

Interference

In the preceding chapter the different types of transmission were discussed in relation to bandwidth. It should be apparent that analogue systems in general require greater bandwidth than digital and hence the receiver becomes less selective.

Interference falls into several different categories according to its nature, but for the present only the types that cause modification to the control signals in the receiver will be considered. Analogue receivers incorporate filter networks to separate the channels. If the interference has inserted signals, the frequency of which falls outside the pass band of the filter, then only the correct signal will be passed. Since the following circuits average out the signals passed by the filter, any interference that causes either extra pulses or gaps will be included in the averaging process. The amount of interference that can be tolerated is therefore dependent upon the degree of control that it is essential to maintain.

The digital receiver passes its signal to the decoder which is stepped on by the receiver pulses. Thus extra pulses or missing pulses result in the stepping sequence synchronisation being lost. If the interference is brief, synchronisation will be restored on the next information frame. However if the interference causes extra pulses, and persists, the synchronisation pulse will contain pulses and the decoder will not be able to lock into the signal.

Digital systems therefore demand a greater performance of the communication link than is necessary for analogue, but when considering the latter it is essential to ensure that the inherent advantages of its transmission are not disguising an inferior receiver design.

Failsafe

Analogue systems more readily lend themselves to the incorporation of failsafe circuits, and generally these are essential to avoid the servos moving to full deflection when the transmitter is switched off. Since failsafe is a dubious requirement, in the case of an aircraft it will still invariably crash, the incorporation of

such circuits in digital systems is rare. In the event of the transmitter signal being lost, with a digital system the servos will not receive any command, and so will remain approximately at their last command position. This is unlikely to result in a response from the model that is any less predictable than with failsafe operation. The inclusion of circuits to provide failsafe probably achieves nothing except a decrease in reliability due to the risk of a failure in one of the extra components.

Inherent reliability

The reliability of an electronic circuit is dependent upon the number of components employed, and the mean time between failures of each different type. If discrete components are used, an analogue system has approximately 2/3 the number of parts to the equivalent digital equipment. Hence it is inherently more reliable. With the adoption of integrated circuits the balance swings rapidly in the favour of digital, mainly due to the reduction in the number of soldered joints.

Cost

With the lower component content of analogue systems their cost is theoretically cheaper. However if a four channel system is considered, the major proportion of the cost is in the mechanical components and these are not affected by the type of system. Thus in practice there is no real price advantage of one type of system over the other.

Future trends

With the commercial considerations of reducing price both by cutting component costs and easing production methods, a trend towards the use of integrated circuits is becoming evident. This favours the use of repetitive types of circuits such as are found in digital systems. Some analogue equipment does have limited application on these lines but unless a similar degree of miniaturisation can be achieved for the remaining circuitry, the long term success of these systems is somewhat dubious in a commercial environment.

The intention is not to belittle present day analogue systems, far from it, they have come a long way towards eliminating the poor reputation caused by indifferent performance of some of the early systems. However the advantage is heavily in favour of digital from a commercial viewpoint, due mainly to the wide range of digital components readily available from other branches of the electronics industry. Because of the predominance of digital, the remainder of this book will be devoted to this type of equipment. However, the sections on superhet receivers and transmitter output stages are, of course, equally applicable to analogue systems.

CHAPTER 5

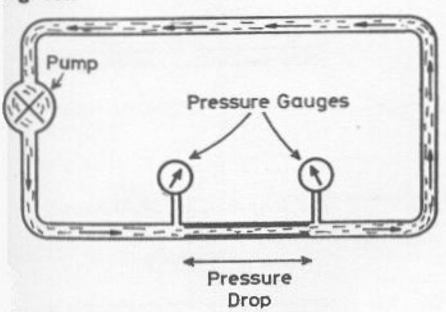
CIRCUIT COMPONENTS AND APPLICATIONS

IN this chapter the aim is to give those readers who do not have a knowledge of electronics, sufficient understanding of various components, that they may appreciate the descriptions of circuits in later sections. A little basic electronics is also included to give a reasonably complete picture. Electronic circuits can in many ways be likened to hydraulic systems, and since the majority of readers will be reasonably appreciative of their features, the analogy will be used wherever it is of assistance.

Current, charge and voltage

An electric current passes through a circuit in a similar manner to the flow of water in a pipe. It must be stressed that current is equivalent to the flow and not to the water itself, since water may still be present when there is zero flow. Water can be stored in containers but flow cannot. Similarly electric current cannot be stored.

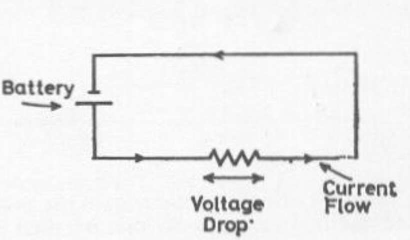
Fig. 18a



Electric charge is the equivalent of water. It can be stored in capacitors, and caused to flow through a circuit. Thus an electric current is the flow of charge through a circuit. In order to cause water to flow through a pipe, a pressure differential must exist between its ends. Similarly for charge to move through a circuit, a voltage difference must exist across the circuit. In Fig. 18, the electric circuit is represented by the resistor and the wires by low resistance connections. The hydraulic system comprises a narrow bore pipe causing a restriction, and large bore, low restriction connections.

If the pump is operated, a pressure differential is developed across it, which causes the water to flow around the system. Two pressure gauges inserted in the positions shown will give different readings. This is because the restriction causes a pressure drop. In the electric circuit the battery is equivalent to the pump. A voltage exists across it due to the internal reaction, and this causes charge to flow through the circuit establishing a current. In a similar way to the pressure drop that was established across the narrow bore

b



tube, a voltage drop will exist across the resistor. If the resistance of the connections is low compared with that of the resistor, the voltage across the resistor is that of the battery.

Ohm's Law and resistance

A relationship exists between the current through a resistance and the voltage across it, and this is defined by Ohm's Law. This states that the voltage across a resistance is directly proportional to the current through it, and the units are so chosen to make the constant of proportional unity.

i.e. Volts (V) = Amps (A) × Resistance (R)

This relationship can be used to determine the effect of resistors connected in series and in parallel. Fig. 19 shows a series combination. Clearly the same current flows through both R_1 and R_2.

Thus $V_1 = IR_1$ and $V_2 = IR_2$

Since $V = V_1 + V_2 = IR_1 + IR_2$
then the effective resistance $R = R_1 + R_2$

For the parallel combination shown in Fig. 20 the result is not so obvious.

Since the same voltage exists across both resistors

$$V = I_1R_1 = I_2R_2$$

Also $I = I_1 + I_2$

Applying Ohm's Law. $V = IR$ where R is the effective resistance

$$\therefore I_1R_1 = (I_1 + I_2)\,R$$

But from above

$$I_2 = \frac{I_1R_1}{R_2}$$

$$\therefore \quad I_1R_1 = \left(I_1 + \frac{I_1R_1}{R_2}\right) R$$

$$\therefore \quad R = \left(1 + \frac{R_1}{R_2}\right) R$$

Hence

$$R = \frac{R_1R_2}{R_1+R_2}$$

Thus the effective resistance is the product of the two resistors divided by their sum.

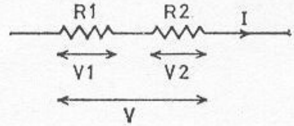


Fig. 19

Capacitors

If we continue with our hydraulics analogy, a capacitor may be likened to a balloon reservoir connected into a pipe, as shown in Fig. 21. If the balloon contains no air prior to the flow being established, it will fill with water as the pressure increases. The degree of inflation will be

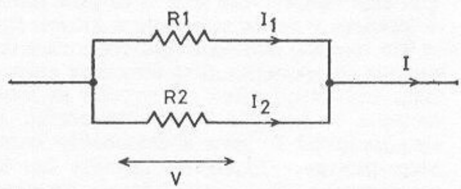
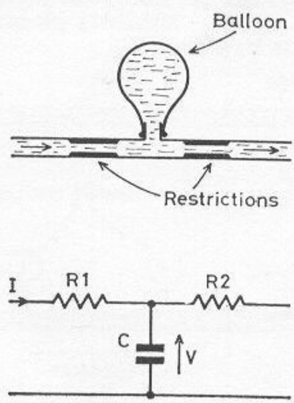


Fig. 20

dependent upon the pressure at the point where it is tapped into the pipe. Thus the balloon stores a quantity of water dependent upon the pressure. If the pressure which established the water flow is removed, the time taken for the balloon to deflate will be dependent upon the degree of restriction in the pipes, and the amount of water stored in the balloon.

A capacitor acts in a similar way to

Fig. 21



store charge. Whilst current is flowing in the circuit, a quantity of charge will be stored, dependent upon the voltage at the point of connection, and the capacity of the capacitor. When the current is removed from the circuit, the voltage across the capacitor decreases as the charge flows away through the resistors in the external circuit. The time taken is dependent upon the size of the capacitor and the resistance. A typical capacitor discharge curve is shown in Fig. 22. The time constant of a resistor network is defined by $C \times R$ where C is in Farads, and R in Ohms. The mathematical law which defines the shape of the curve is such that the voltage falls to half of its previous value in a time of 0.69CR. In digital equipment considerable use is made of this in the various timing applications.

Inductors

An inductor is a device which tends to oppose any change in the current flowing through it. This is done by storing energy in a magnetic field. If the current through an inductor decreases, the stored energy causes a voltage to be induced across the windings, and this tends to maintain the circuit in its previous condition. Thus an inductor has two important properties. A current flowing through it causes a magnetic field to be produced, and a changing magnetic field causes an induced voltage. Both these properties are used in a transformer.

If two coils are arranged in such a way that they are linked by the magnetic field of each other, a change in the current flowing in one coil changes its magnetic field, and the second coil is acted upon by this change. The result is that a voltage is induced across the second coil, and a current can then flow in a circuit connected across it.

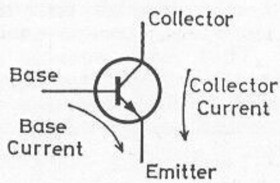


Fig. 23

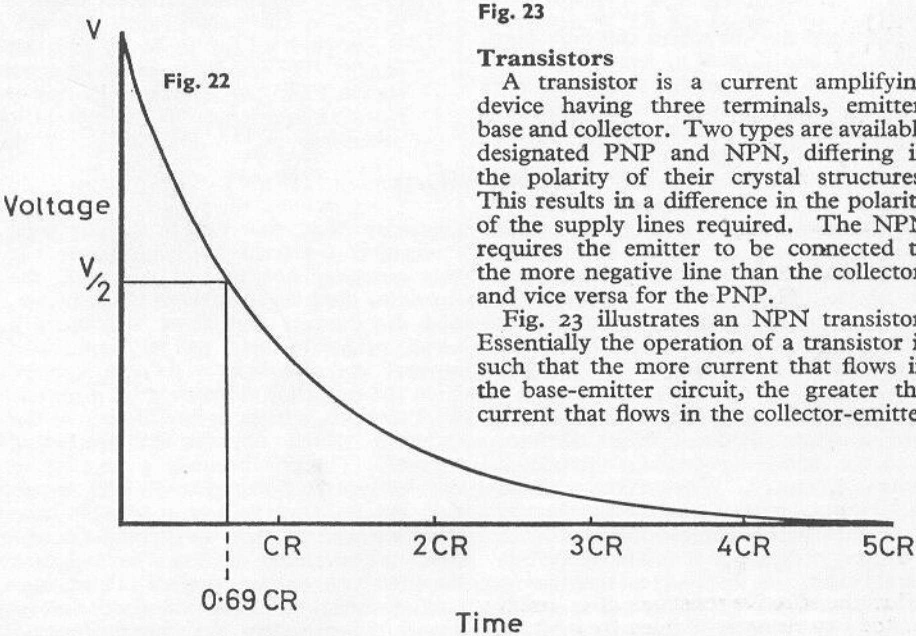


Fig. 22

Transistors

A transistor is a current amplifying device having three terminals, emitter, base and collector. Two types are available designated PNP and NPN, differing in the polarity of their crystal structures. This results in a difference in the polarity of the supply lines required. The NPN requires the emitter to be connected to the more negative line than the collector, and vice versa for the PNP.

Fig. 23 illustrates an NPN transistor. Essentially the operation of a transistor is such that the more current that flows in the base-emitter circuit, the greater the current that flows in the collector-emitter

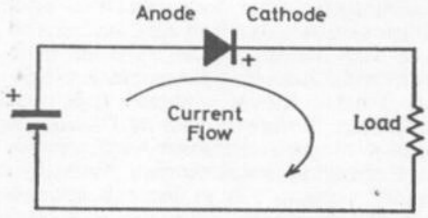


Fig. 24

circuit. Very approximately the ratio of base current to collector current is the d.c. gain of the transistor. Consider a transistor with a gain of 20. If a current of 0.5mA flows in the base circuit, 10mA can be taken by the collector circuit. If the base current is increased to 1mA, the collector current increases to 20mA. By causing these currents to flow through resistors, the amplified signal may be converted to a voltage. Thus with suitable circuit configurations a small voltage signal can be supplied to a simple transistor circuit, and an amplified voltage signal obtained at its output.

Diodes

Diodes are devices which can only pass current in one direction, from anode to cathode. All the time that the anode has a more positive voltage on it than the cathode, the diode will conduct. If the anode goes more negative, then the doide is said to be reverse biased, and no current can flow. A little confusion can occur over the direction in which the diode should be connected. This is because the cathode is usually referred to as the positive side of the diode. Fig. 24 shows the reasoning for this. The direction that current can flow is such that the cathode is the positive connection to the external load circuit.

There exists a special type of diode called a zener diode. When forward biased, i.e. positive potential to anode, it conducts normally. When reverse biased it exhibits a special characteristic whereby at low voltage no current flows as with an ordinary diode, but when a certain critical voltage is applied current flows. In this condition, the voltage drop across the diode is virtually independent of the reverse current flowing through it, the drop being determined by the nature of the crystal structure in the diode. Used in this way, a zener diode can provide constant voltages for reference and regulation purposes.

After this brief summary of the functioning of individual components, their applications in some simple circuits will now be considered.

Tuned Circuits

The heart of any superhet radio control receiver is its I.F. (intermediate frequency) strip, and the R.F. (Radio Frequency) tuned circuit. In both cases circuits employing inductors and capacitors in parallel are used to tune the receiver stages to particular frequencies.

Fig. 25 shows a basic R.F. tuning circuit.

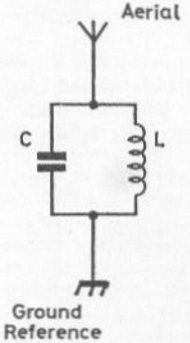


Fig. 25

Consider that the capacitor has been charged to a voltage by some means. If this charging potential is removed, the capacitor discharges through the inductor, and the current that flows will cause a small magnetic field to be established around the inductor. As the current from the capacitor dies away, the magnetic field causes a voltage to develop across the inductor which opposes the decreasing current. There is however a time lag in the voltage appearing, with the result that the induced voltage is greater than that required to just nullify the current flow. The result is that the capacitor becomes charged up again. There is a similar time lag in the voltage build up across the capacitor due to the current

flowing in, so that the capacitor voltage tends to rise as the induced voltage decreases. This brings us back to the starting condition so the sequence of events commences again.

If no energy were lost, this system would be in continuous oscillation at a frequency dependent upon the values of the capacitor and inductor. However an inductor will have a small value of resistance and so when current flows through it, heat is generated and energy is lost as a consequence. If the oscillations are to be maintained energy must be supplied externally. In the R.F. stage an aerial is connected to the system and "collects" a small proportion of the energy radiated from the transmitter. Provided the circuit is tuned to the same frequency as the incoming signal, the received energy will always be supplied to the inductor-capacitor circuit at the same stage of the cycle of oscillation. If sufficient energy is supplied, the losses in the system will be overcome and oscillation will continue. Thus, provided the transmitter signal is sufficiently strong, the R.F. receiving circuit will be maintained in oscillation.

In the I.F. strip, the situation is very similar. The tuned circuit in this case, is the primary winding of the I.F. transformer and its capacitor, and the external energy is supplied by the transistor stage driving it. (Fig. 26). The secondary

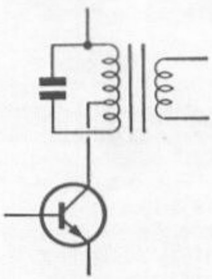


Fig. 26

winding is linked with the changing magnetic field of the primary by the ferrite former on which they are both wound, and the result is that a voltage is induced in this secondary winding. It is beyond the scope of this book to go into transformer theory, but suffice it to say that the turns ratios are chosen to match the external circuits to the oscillating circuit. It is for this reason that the primary circuit is fitted with a tapping point.

A similar case exists with double tuned circuits, the difference being that both primary and secondary windings are tuned to the resonant frequency. For a given level of oscillation in the primary, a larger output can be obtained if the secondary is also tuned. A signal of slightly incorrect frequency in the primary produces a lower level of oscillation, and for a tuned secondary winding, its response will also be less. Without the tuning, the signal level in the secondary would be almost a linear relationship with that of the primary. Thus a double tuned circuit provides a greater degree of signal selectivity.

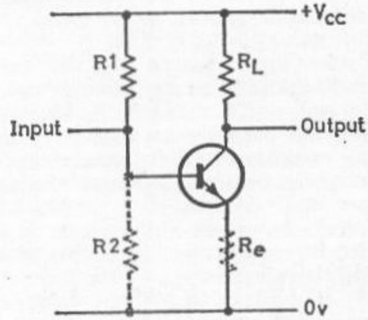


Fig. 27

Transistor Amplifier Circuits

Consider the simple circuit shown in Fig. 27 which is required to amplify a sinusoidal input signal without distortion. Since a sine wave goes both positive and negative about its reference level, the output point (the collector of the transistor) must be biased so as to permit a swing in either direction. For a supply voltage of +Vcc, to permit the maximum possible swing in both directions, the output must be biased to $+\frac{1}{2}$Vcc in the quiescent state. The output signal can then swing approximately between +Vcc and 0v. With a load resistor R_L, the biasing could be achieved by resistor R_1 alone. The value of R_L and the voltage $+\frac{1}{2}$Vcc define the current that is required

to flow in the collector circuit, and then dividing this by the transistor current gain, the base current is calculated. This in turn defines the value of R_1. This is not a very satisfactory method of biasing since any variation in transistor gain will change the bias level. To overcome this, resistor R_e is included. Since the collector current flows through this resistor, a voltage drop occurs across it. The base current is dependent upon the value of R_1, and the voltage difference between the supply line and the voltage at the emitter of the transistor. Thus a high gain transistor will tend to produce more current in the collector circuit and hence a greater voltage drop across R_e. This reduces the voltage across R_1 so reducing the base current. This in turn controls the collector current so the overall result is that the biasing level becomes a function of the resistor values rather than the transistor gain.

A further improvement may be made by including R_2. With the biasing system outlined above, the value of R_e becomes large, and the base current small. Temperature variations can produce quite large changes in small currents through semiconductor devices so a method is introduced to increase the current in R_1 in order to minimize the effects of changing currents. R_1 is therefore decreased in value with a consequent increase in its current, but to prevent this flowing through the transistor base R_2 is added.

The overall gain of the stage when amplifying our sinusoidal signal now becomes a function mainly of R_1, R_2 and R_e. This is very desirable since if the transistor gain had a large influence on the stage gain, different examples of the same circuit would begin to distort the output signal, at widely differing levels of input.

Circuits of this type are to be found in the I.F. strip of a superhet receiver, the only real difference being that R_L is replaced by the primary winding of an I.F. transformer. The bias level for transformer coupling is subject to different considerations for distortion, but the principle is the same.

Where an amplifier stage is required to handle pulses, the input will swing in one direction only from the quiescent condition. Referring to Fig. 27 again, if the circuit were required to accept negative going pulses, R_2 and R_e are not required. The value of R_1 is chosen so that with the minimum transistor gain, sufficient base current flows to produce a collector current that will cause a voltage drop across R_L that is equal to Vcc. If more base current is caused to flow, the collector current cannot be increased due to the limiting effect of R_L. The transistor is then said to be in a state of saturation.

If a negative going pulse is applied to the transistor base, the base current is reduced for the duration of the pulse. The collector current decreases causing less current to be taken through R_L with a resultant decrease in the voltage drop. The voltage at the collector therefore increases. Hence the negative going pulse on the base has produced a positive going pulse at the collector. This inversion of the signal is an important property of single transistor amplifier stages.

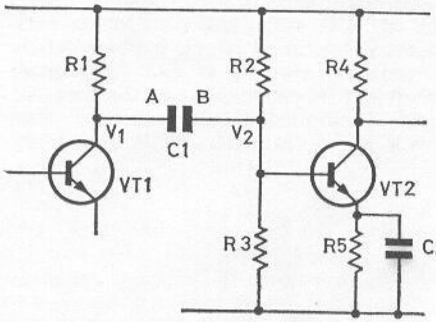


Fig. 28

Capacitive Coupling

In the preceeding section we have seen how the biasing conditions for an amplifier stage may be set up. If two such stages are to be connected with the output of the first acting as the input of the second, a resistive connection cannot be used since this would interfere with the d.c. conditions. One method of overcoming this is to use capacitive coupling as shown in Fig. 28. If no signal is present a d.c. bias of V_2—V_1 exists across the capacitor.

If the signal output of the first stage produces a transition of v at the collector, say in a negative going direction, plate A of the capacitor will have a voltage of $V_1 - v$ applied to it. Now the effective input resistance of the second stage, made up of the combined effects of R2, R3 and VT2, will be relatively high. Therefore the time required for current to flow into the capacitor and cause a new distribution of charge, will be long compared with the transition on plate A. Plate B will therefore tend to follow the transition applied to plate A and hence the input of stage 2 will see the transition. If the output of stage 1 quickly returns to its quiescent state, i.e. before the new charge distribution has had time to become established, plate B will also return to its original voltage. The second amplifier stage has therefore had the complete signal applied to its input.

The value chosen for the coupling capacitor is dependent upon the frequencies of the signal that it is to handle. Its value must be sufficiently large so that in the duration of the signal, the charge in the capacitor does not change significantly. If the capacitor is too small, the second stage will receive a negative "pip" as the first stage output goes negative, and a positive "pip" as it returns to the original stage. This is because new charge conditions will have become established in the capacitor, before the trailing edge of the signal goes positive.

Decoupling

In many applications there will be parts of circuits which should remain at a relatively fixed voltage level, but which for some reason or other exhibit a tendency to change when a signal is being handled. A typical point with this tendency is the emitter of transistor VT2 in Fig. 28. When a signal is applied to this circuit, the transient produced in the collector current also occurs in the emitter circuit. The result is that a voltage change occurs at the emitter, and in the opposite sense to that at the collector. Generally this is undesirable since it has the effect of reducing the signal gain of the amplifier stage. To overcome this a large value capacitor is connected from the emitter to the 0v supply line as shown. When the emitter voltage begins to change with the effect of the signal, the voltage at the emitter can only change significantly, if the duration of the transient is sufficient for the capacitor to charge or discharge through the effective circuit resistance, With a large value capacitor, the emitter is therefore held at a fairly constant voltage.

In certain cases where the gain of a stage is deliberately required to be restricted, the decoupling capacitor may be omitted.

Another common application for decoupling is on supply lines. Since these can be carrying quite large currents, even a very low resistance can cause voltage drops. If the currents switch on and off (e.g. when a servo operates), the effect is to cause pulses to appear on the supply lines. A large value capacitor across the supply lines, at the opposite end to the battery, minimises the effect of these switching transients. The capacitor has to be a much larger value than for other decoupling applications since the resistance with which it is acting is of a much lower value.

It will be observed in some circuits where high frequencies are present, that a small value capacitor is connected in parallel with the large one. This is necessary since the physical construction of large value electrolytic capacitors, results in an effective resistance being introduced within this type of capacitor. Thus for high frequency transients, the electrolytic capacitor is not very effective. However since the value of capacitance required to suppress high frequency transients is less, capacitors of a different type of construction can be used effectively. Thus by putting a disc ceramic capacitor across an electrolytic, both high and low frequency decoupling is achieved.

Diode Gates

In any system where information on a single highway represents a composite function, circuits have to be employed to effect the combination. In digital logic systems, such circuits are termed gates, and these take one of two basic forms. An

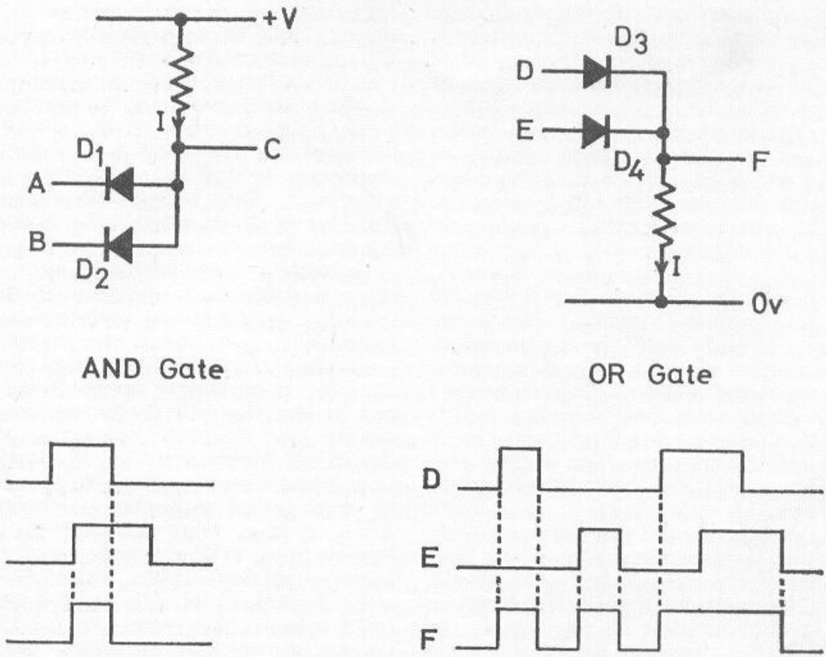


Fig. 29

AND gate gives an output only when all of its inputs are present, and an OR gate gives an output when one or more of its inputs are present. Fig. 29 illustrates these together with their appropriate signal waveforms. A signal is present when the waveform is positive and no signal is represented by ov.

For the AND gate, when neither of the inputs is applied current flows from +V through the resistor and both diodes into the circuits which switch the inputs. This causes a voltage drop across the resistor and provided sufficient current flows, the output at C will be nearly ov. If one input goes positive, say input A, current still flows through the resistor and D2, although D1 is reverse biased and therefore not conducting. Output C therefore remains unchanged. If both A and B are positive, both diodes become reverse biased and no current flows. Output C therefore rises towards +V. If either or both inputs return to ov, output C also falls. Thus the output is positive only whilst both inputs are positive.

The OR gate functions similarly, except current enters from the switching circuits driving the inputs, and flows to the ov line. If both inputs are at ov or slightly negative, neither diode can conduct, so no current flows through the resistor. Output F is therefore ov. If the input to either diode goes positive, the diode becomes forward biased and current flows. Output F therefore also goes positive. If both inputs are positive the output will also be positive, an ov output only being obtained when both inputs are low.

An OR gate of this type is usually found in digital transmitters for combining the pulses from individual channels before applying them to the R.F. circuits.

In any application of diode gates consideration has to be given to the thresholds at which the gates begin to change state. This is because for a typical silicon diode when conducting, approximately ½ volt is developed across it. In

the case of the AND gate, this means that the output (Point C in Fig. 29) cannot be less than ½ volt in the no output conditon. Simarlarly the OR gate output will be ½ volt less than the highest voltage applied to any of its inputs.

Both types of gates can be extended to function with more than two inputs simply by adding more diodes. A difficulty that could be experienced might be variations in the output driving capability dependent upon the number of inputs that were in the on or off state. To overcome this a switching circuit is usually connected at the output of the actual gate. If this switch inverts the signal, as in the case of a single transistor, then the gate becomes a NOT AND or a NOT OR type. These are usually referred to as NAND and NOR respectively. The names are derived logically, since if we consider the output of the AND or OR gate to be true (i.e. the input conditions are satisfied), when the output is high, then it is not true when the output is low. Hence an inverting amplifier on the output gives a high signal when the input conditions are not satisfied, and a low signal when they are correctly conditioned.

The reader might like to work out how an OR function can be derived from an AND gate with inverting amplifiers on both the input and output. The chapter on decoders includes examples of the type of approach that is usually adopted in solving logical system design problems.

Integrated Circuits

In computer systems a large number of identical circuits are used due to the repetitive nature of the logic design. Consequently, some years ago computer manufacturers began to demand semiconductor devices that could perform more complex functions than could be achieved by a single transistor. Transistor manufacturing had reached a stage where they were produced by depositing some form of crystal or oxide onto a base slice of another crystal, and then etching out patterns which had been reproduced photographically. It is an obvious extension of this technique to produce more than one transistor on a single crystal slice, and if appropriate interconnections are included, the device can operate as a particular circuit. Diodes can be produced as effectively as transistors, and consequently diode gates with transistor amplifier stages on their outputs can be produced with relative ease. Other basic functions encountered in logic design can be produced by appropriate interconnection of gating circuits, and so a natural extension has been for semiconductor manufacturers to produce a range of integrated circuits covering the basic functions.

Since the initial development and tooling cost is somewhat high in semiconductor manufacture, it is not unexpected to find that the devices readily available are aimed at the area of the market that is likely to provide the largest sales, i.e. the computer market. Also digital circuits only require an output to be either on or off as the final transistor stages switch. What happens in between is of little consequence. This makes the design of integrated circuits for digital circuits far less critical than for say an ordinary amplifier. However with the experience gained in manufacturing digital circuits, techniques have improved to give greater consistency, with the result that a large variety of so called linear circuit functions (i.e. those that are not digital in nature) are becoming readily available. Improved manufacturing techniques have also allowed complex digital functions to be produced on single semiconductor chips only about 4 millimetres square. Connections to chips of this size obviously cannot be made easily so the manufacturers bond leads onto the semiconductor and encapsulate the whole assembly.

The most common shape of encapsulation is the dual in line (D.I.L.) package with either 14 or 16 pins arranged in two rows, although some simple functions requiring less leads can be supplied in an encapsulation that resembles a multi-lead transistor. For military applications where space is often at a premium, another type of package known as the " Flat Pack " is used. This is smaller than the dual in line type and generally more expensive, hence it is not often found in radio control equipment.

The great advantage of integrated circuits, apart from their miniaturisation, is a

large increase in reliability, compared with similar circuits assembled from discrete components. In the case of complex functions, from a manufacturing viewpoint, the time saved in assembly can represent a considerable cost saving.

The application of integrated circuits to radio control has been somewhat slow due to the limited areas in which they could be applied

In a digital transmitter where the timing circuits usually consist of a single transistor circuit for each channel, and high stability components are essential, the use of integrated circuits is almost completely uneconomic. It is possible that in time, a circuit will become available at the right price, which can perform this function. The pulses from each channel have to be combined with an OR gate, and some designers are using integrated circuits here because it is possible to achieve a degree of pulse shaping with relative ease. However, since integrated circuits generally operate on a 5 volt supply, difficulties are encountered with the higher voltages usually employed in the transmitter.

The receiving system does not suffer from the operating voltage anomaly of the transmitter, since a 4.8 volt supply is normally used. Standard digital logic circuits can be employed in the decoder, and with careful design some of the latest circuits introduced for commercial radio applications, can be used in the R.F. and I.F. areas.

A typical servo amplifier employs circuits that are partly digital and partly linear in nature. First attempts to use integrated circuits for this purpose centred around several transistors in a single encapsulation used in fairly conventional circuits. A number of semi-conductor manufacturers now produce integrated circuits especially designed for servo applications. A single encapsulation includes all the circuit components except those for the timing and feedback and perhaps the output transistors. Such circuits offer improved performance over discrete types since increased complexity is possible without an increase in physical size. For a commercial manufacturer there is the advantage of reduced assembly and test time, to offset the cost of the integrated circuit.

Standard TTL (transistor-transistor logic) integrated circuits have a comparatively high current consumption and they can only offer a real advantage when a wide area of the system can be fulfilled by their use, e.g. a decoder. The introduction of low power TTL would have overcome this problem when prices reached a competitive level, but this has been over-shadowed by the introduction of CMOS (Complementary Metal Oxide Semi-conductor). Integrated circuits manufactured by this technique have reduced current consumption in most applications and will operate on a wide range of supply voltages. Hence they are now finding usage in both transmitter encoders and channel decoders. The following chapters on the various units which make up a radio control system will discuss the use of integrated circuits in more detail.

CHAPTER 6

DIGITAL TRANSMITTERS

THE nature of the signal employed to transmit control information in digital systems was discussed in Chapter 3. It was shown that a required servo position is defined by the width of a pulse, and that the information for several servos is sent sequentially followed a synchronisation pulse. One complete set of pulses is termed a frame of transmitted data, and provided that this frame is repeated sufficiently fast, the time delay in a servo responding to a change in its demanded position, is negligible. In practice frame rates of around 50 to 60 frames per second are employed, and since the mechanical inertia of a servo restricts its response speed, there is no sense of discontinuity in its performance due to time multiplexing of the channel information.

Synchronisation

With a time multiplexed signal, the receiver is required to scan the incoming information and separate each channel onto different outputs. To do this a synchronising command has to be given so that the information from one channel does not appear on the output of another. The specification of the synchronisation requirements is dependent upon what degree of error can be tolerated when using the system. By choosing as high a frame rate as possible, the effect of loss of synchronisation during a single frame can be made negligible, provided resynchronising occurs within the next few frames. A system could be designed which sent several frames of channel information between each synchronisation pulse, and this would have the advantage of greater control accuracy due to the higher repetitive rate achieved. However other considerations restrict the frame rate so that the number of frames that could be transmitted between each synchronisation period, is somewhat small. Hence the extra complication in circuitry is not justified and a slower rate of information transfer is accepted together with the inclusion of a synchronisation pulse within each frame.

The synchronisation pulse has to differ in some way from the command pulses so that it can be readily detected by a circuit in the decoder. It can be made either longer than the longest channel command, or shorter than the shortest. The lätter has the advantage of allowing a higher information transfer rate but the difficulty of detection in the receiver does not justify its use. The synchronising pulse therefore takes the form of a pulse of at least twice the length of the maximum channel command pulse, and this can be detected by a simple resistor-capacitor network. This will be considered more fully in Chapter 8.

The Modulating Signal

Since the channel pulses are to be transmitted in sequence, the simplest way of achieving this is to generate them sequentially. Thus the start of the generation of a channel pulse is initiated by the trailing edge of the preceding pulse, and the first channel pulse is initiated by the circuit controlling the frame rate. This is shown in Fig. 30. The modulating signal is only required to specify the time at which the transitions

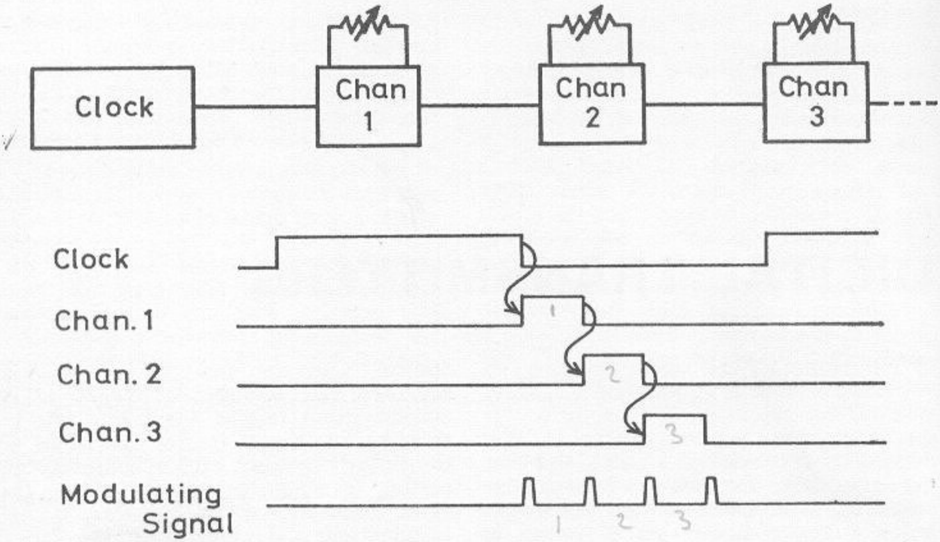


Fig. 30

occur for the full control information to be defined. This is termed differential encoding, and the time period between two successive pulses defines the width of the channel command pulse.

If a control pulse increases in width since its trailing edge occurs later, the start of the next channel pulse is delayed. Thus the trailing edge of the last channel pulse occurs later. If the clock is of a fixed period, the time between the trailing edge of the last channel pulse and the beginning of the next frame may be used for synchronisation. Fig. 31 illustrates this for a two channel system and shows how the free running clock frequency must be chosen to allow the full time period for synchronisation to be achieved, irrespective of the command pulse widths.

The time period A when the controls are at a minimum position represents wasted time since no useful information is transferred. For a system employing several channels, say four or more, this period could typically be as long as 4 milli-seconds, for a clock cycle of 16 milli-seconds. This is clearly inefficient, and since servo resolution is somewhat dependent upon repetition rate, it is undesirable. An improvement to this system is to use a fixed length synchronisation pulse and absorb the changing channel commands by varying the frame rate. This can readily be achieved by using one portion of the clock cycle to generate the synchronisation, and overriding the other portion of the cycle with a signal derived from the trailing edge of the last channel pulse. This is illustrated in Fig. 32, and in terms of circuit design is easy to achieve.

The choice of the modulation pulse width and its variation with the commands is dependent upon many factors which are determined by the overall system. It has already been shown that short pulses are transmitted which define the change from one channel to another, and the minimum width of these is a function of the receiver bandwidth. A typical narrow bandwidth receiver will have a useful bandwidth of around 6KHz, although it may respond in some manner to signals on either side of this. The rise and fall times of the transmitted pulse must therefore be chosen so that the receiver is able to accept them without excessive degradation. If the rise and fall times are made too fast, there is a poss-

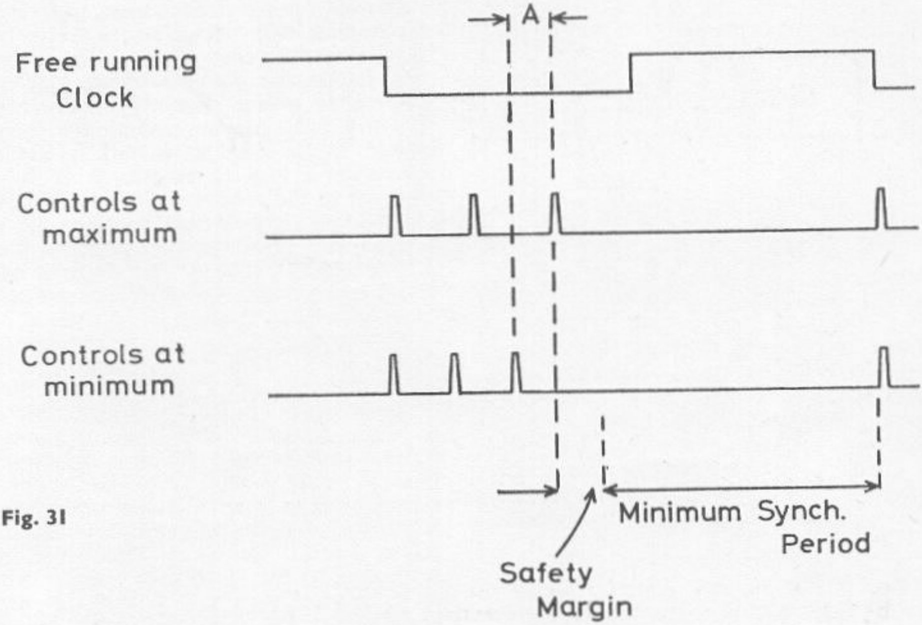
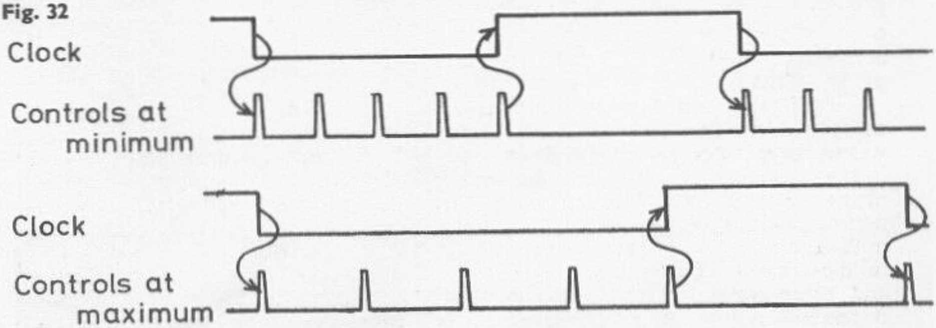


Fig. 31

ibility of spurious oscillations being produced in the receiver I.F. strip.

In Chapter 3 the effect of passing square waves through filters was discussed, and the same principles apply in this instance. If we wish to choose a signal that can be accepted by the 6KHz bandwidth receiver, we design for 5KHz in order to allow a margin of safefy. Referring back to Figs. 11 and 12 we find that a square wave of 5KHz repetition rate could be accepted, i.e. the periodic time of the pulse is 0.2mS. During this time period, the square wave has to start from its mean level, rise to a maximum, fall through the mean to its minimum level, and then rise again to the mean. Fig. 33 illustrates this. Now we are only interested in one transition from the mean and if the sine wave is to be passed without distortion or loss of amplitude, this transition must be completed within 0.05mS. If we consider our requirements a little closer we find that we are not permitting complete oscillations to occur and it is more accurate to consider the

Fig. 32



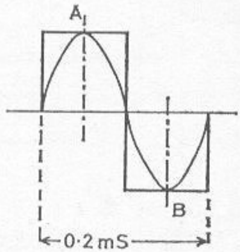


Fig. 33

rise or fall time of our modulation pulse to be determined by the time between A and B, i.e. 0.1mS. This can be shown theoretically to be the case, but is beyond the scope of this book. The outcome of this discussion is that 0.1mS must be allowed for the rise and fall of the modulating signal.

Since a square wave has component frequencies within it, a proportion of the signal will be outside the bandwidth

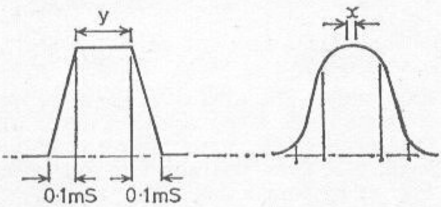


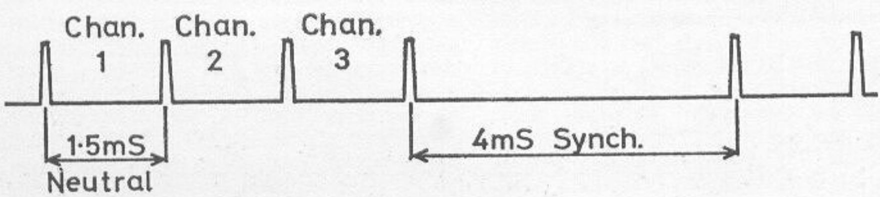
Fig. 34

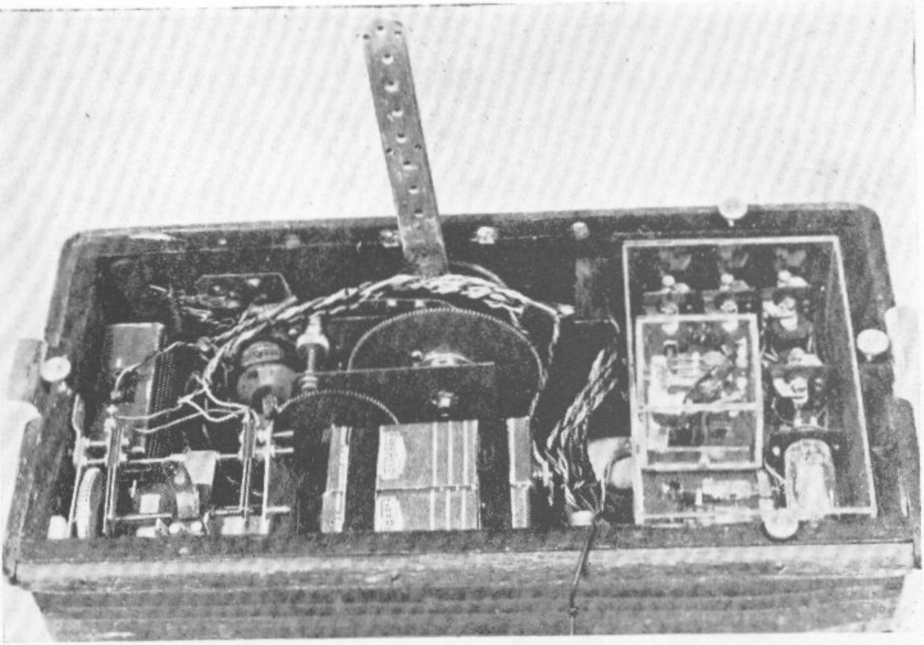
of the receiver and so represents a degree of redundancy. If we control the signal to keep it closer within the bandwidth of the receiver, the signal eventually passed to the decoder will resemble the transmitted signals more closely. Fig. 34 shows how control of the modulating pulse results in the 5KHz bandwidth receiver reproducing the signal with a reasonable degree of accuracy. A time period x is allowed for the disturbances caused by the leading edge transition to settle before the trailing edge occurs. This therefore defines the nominal width y of the transmitted pulse. For a pulse with the rise and fall times as specified y is required to be 0.15 to 0.2mS minimum otherwise a loss of amplitude is liable to occur in the receiver. Our modulating pulse is now completely specified. The rise and fall times must not be faster than 0.1mS and the nominal width at the half amplitude position should not be less than 0.25mS.

The width of the channel command pulses in the minimum position must be such that there is adequate separation between the modulating pulses. A factor of three or four times the modulation pulse width is used giving a minimum channel pulse of 1mS. A shorter channel pulse could be used, but any effects due to the restricted bandwidth of the receiver would introduce significant inaccuracies in the servo positioning. The wide extreme is not limited by any consideration of the transmitter, but is dependent upon the servo amplifier and its ability to resolve changes in channel pulse widths. This will be discussed in detail when servo amplifiers are considered. Suffice it to say that with most types of amplifier circuits, a pulse width variation of 5 microseconds can be resolved. Since this is required to correspond to approximately 0.5% of servo travel, full servo travel can be defined by a 1mS variation. The channel command pulse is therefore made so that 1.5mS corresponds to the neutral position, and the two extremes of servo throw are defined by 1mS and 2mS respectively.

The complete modulating signal for a three channel transmitter is shown in Fig. 35, and the individual pulse in Fig. 36.

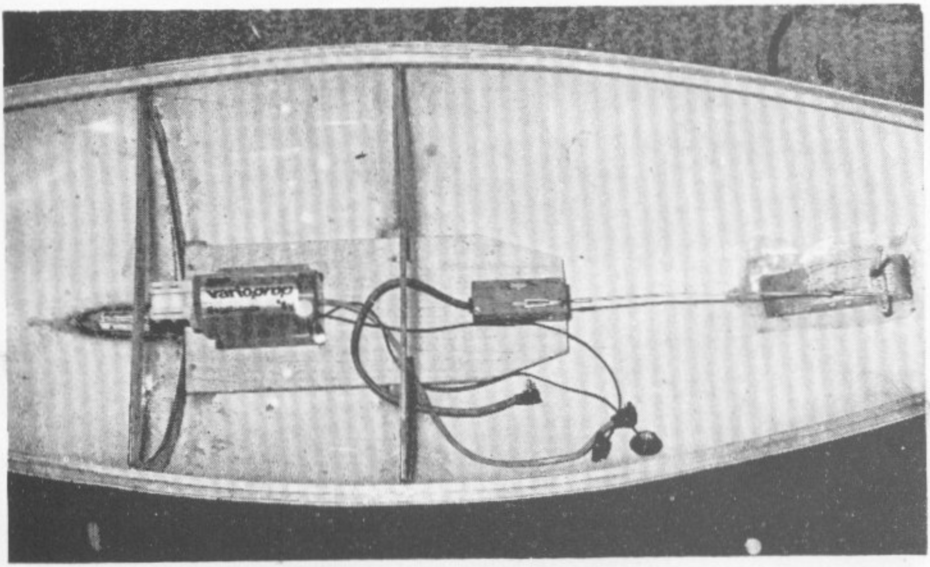
Fig. 35

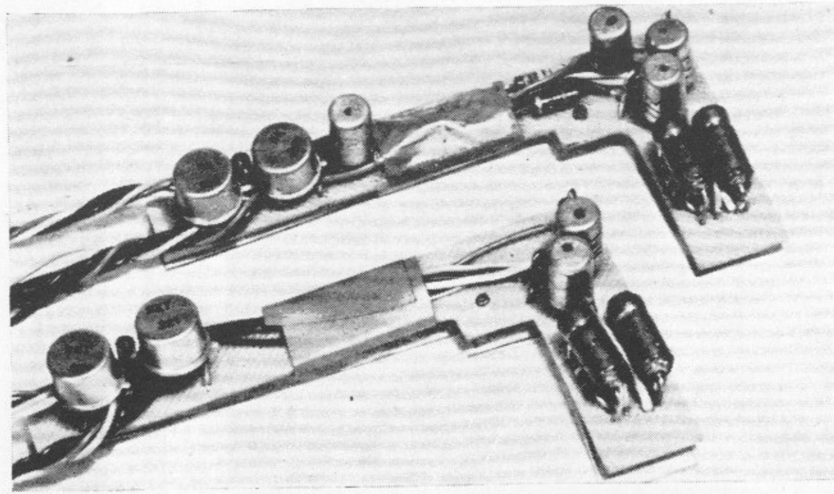




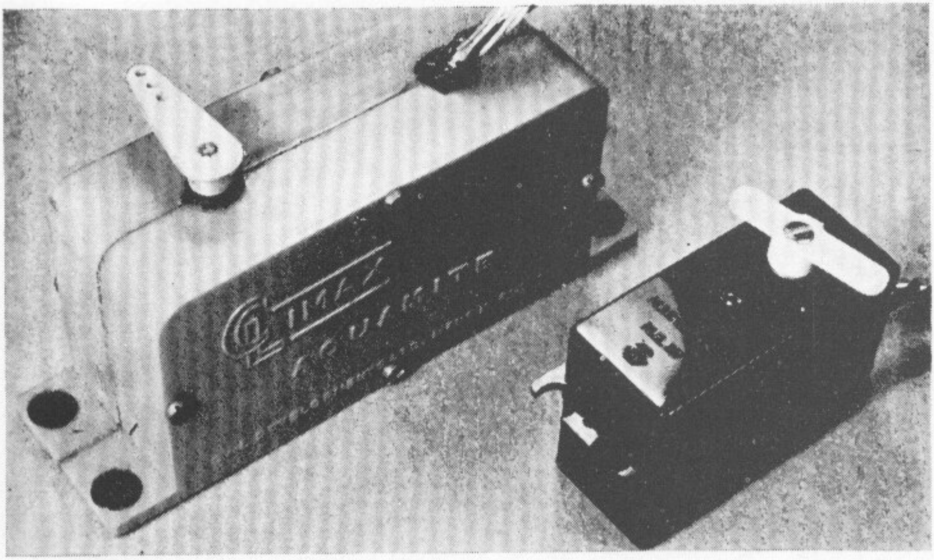
Model yachts require powerful winches and this early example of a complete system in a waterproof box has a 4-channel reed, valve receiver. It is completely home built

Modern solution is simpler—to say nothing of being more compact—a commercial winch and proportional servo. The rest of the equipment is standard aircraft type receiver and transmitter. Winches in both cases are progressive, but proportional would be better

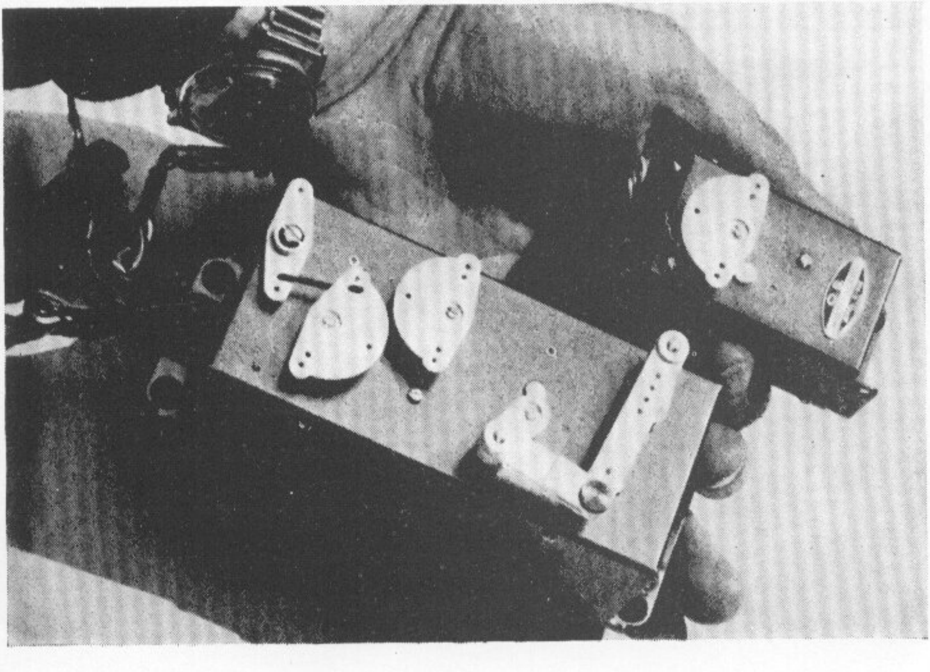


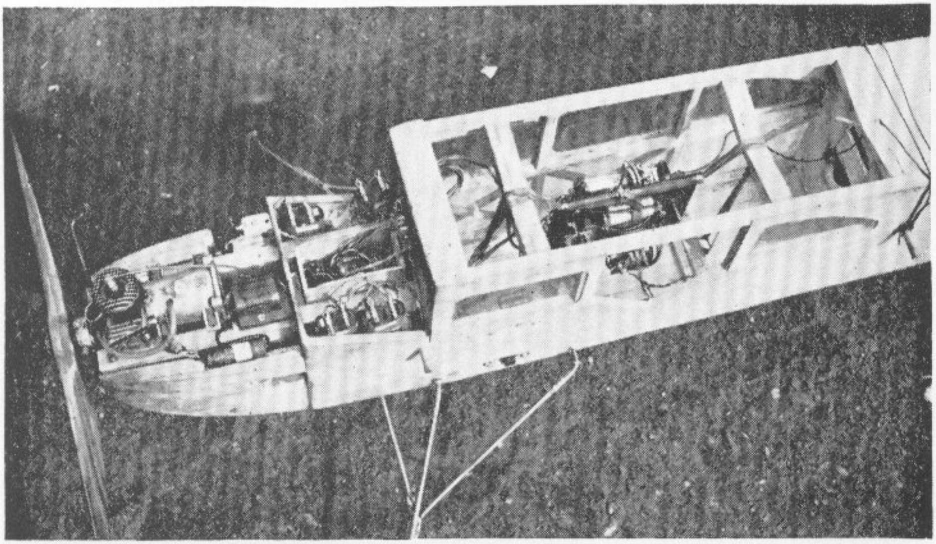


The ubiquitous Bonner Duramite, a really powerful servo for reeds, operated via two relays, but later fitted with an amplifier, the trim and self centring versions of which are shown above. Below are shown two servos set up for elevator, to provide trim and self centring, the trim unit being on the left and the main unit on the right. The amount of trim was variable by repositioning the push rod in the trim bar

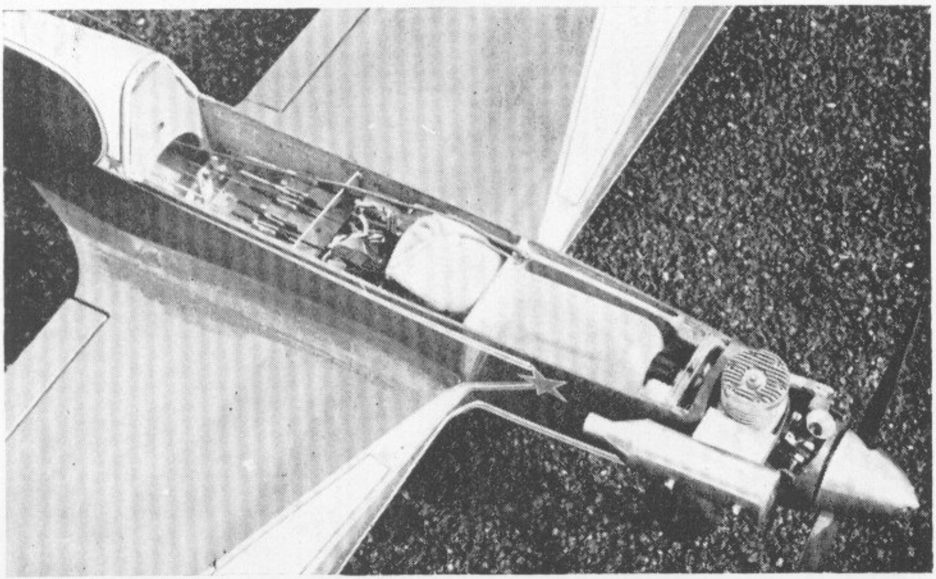


Above: a comparison in size, with the Aquamite, a non-amplified reed servo waterproofed for marine use, and the MacGregor MR-10, a proportional servo typical of the size popularly in use now. Below we have the O.S. reed servo unit. The main case contains the total mechanics and electrics for elevator/rudder/throttle control, while the aileron servo is separate. Note the built-in elevator trim bar and the bellcrank linkage for reversing throttle take-off. Exceptionally well built and reliable, this unit did not " catch-on " because its arrival coincided with the introduction of the first commercial propo outfits





The heavy payload of batteries—not only for the receiver HT and LT and actuator supply, but also the spark ignition engine—gave this single channel model of some 5-6ft. span a rather disproportionate payload. Now compare this photo with a modern aerobatic proportional model, with servos receiver and battery (Nicad) neatly tucked away in a fraction of the space, and contributing only some 20 per cent to the total weight of the model



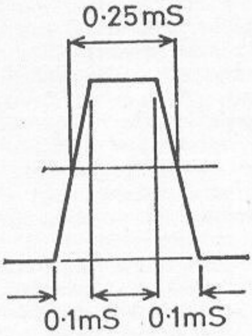


Fig. 36

Clock Generator

Referring back to Fig. 30 it will be seen that the whole of a command pulse chain is initiated by the clock generator. This is a square wave oscillator and usually takes one of two forms, either a unijunction transistor followed by a buffer stage or a multivibrator circuit.

A unijunction transistor differs somewhat in its electrical characteristics from the standard type of transistor, in that it is essentially a triggering device. Fig. 37 shows a simple oscillator circuit together with its associated waveform. Assume initially that the unijunction is not conducting. The capacitor C charges towards +V through resistor R1, and the voltage across it rises. When this voltage reaches approximately 60% of the voltage between b_1 and b_2 (in this case 60% of +V since the unijunction is not conducting and therefore there is no voltage drop across R2), the unijunction starts to conduct sharply and the capacitor discharges through the e — b_2 junction. When the voltage falls below a certain level, the unijunction stops conducting and the capacitor charges up to restart the cycle.

The second circuit is the cross coupled multivibrator shown in Fig. 38. If we consider VT1 to have suddenly started conducting, its collector voltage will have fallen rapidly. The transition is transferred to the base of VT2 via C1 turning it off so that its collector voltage

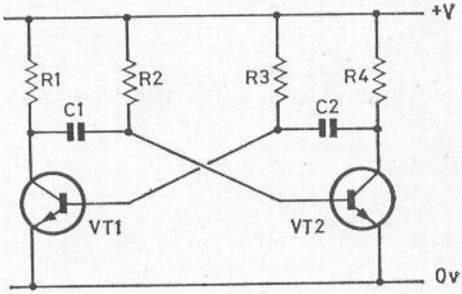


Fig. 38

is high. C1 charges through R2 towards +V until the voltage on the base of VT2 causes VT2 to start conducting, and as a consequence its collector voltage falls. C2 transfers this negative transition to the base of VT1 turning it off and the transistors remain in their respective states until C2 recharges through R3 and

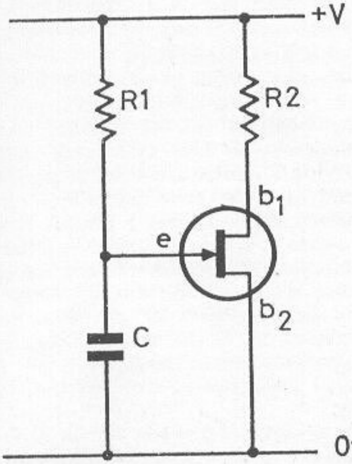
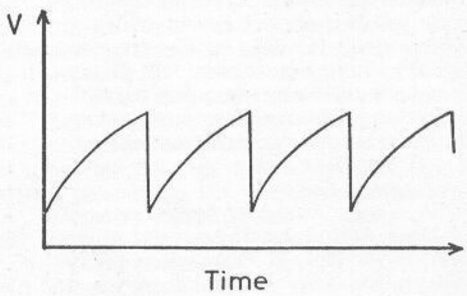


Fig. 37



causes VT1 to turn on again. The sequence then starts again, maintaining a periodic oscillation.

Unlike the unijunction circuit, the multivibrator has two distinct time intervals within one cycle of oscillation. The first interval is determined by the values of R2 and C1 and the second interval by R3 and C2. This becomes significant when a variable frame rate mode of operation is to be employed, since one part of the cycle can be used to completely specify the synchronisation pulse duration. The appropriate resistor can be made variable to simplify the setting up process. The unijunction circuit does not lend itself to the variable frame rate application in the same way.

The advantage offered by the unijunction oscillator is that its frequency is defined by only two components, the capacitor and resistor, whereas the multivibrator requires four. Thus if an accurately defined frequency is required, the unijunction uses only two high stability components together with the higher cost transistor. Unfortunately a buffer stage is required with another transistor, in order to allow coupling in of the first channel control potentiometer. Thus if the overall system is designed so that the clock rate has a wide permissible tolerance, the multivibrator is the most logical choice due to its lower cost.

The degree of accuracy required in the clock frequency is relatively low since only the synchronisation is affected. In the receiver, the pause between frames of informations is invariably detected by a capacitor charging up to a threshold voltage level. Consequently, provided the frame rate always causes a synchronisation pulse duration in excess of the charging time, it does not matter how much the frame rate varies. If the synchronisation detector requires 4mS to operate, a synchronisation pulse of 6mS will be adequate to ensure operation under all conditions of supply voltage and temperature variation.

With a non variable frame rate, the total clock period for the specific figures quoted is the sum of the 6mS synchronisation pulse plus the total of all the channel pulses in their wide extreme positions, i.e. 2mS. Thus for a fine channel transmission the minimum period is 16mS giving a frame rate of 60 frames per second.

If a variable rate is incorporated, the time interval of one half of the multivibrator is set to 6mS, and the other half has a period in excess of the 10mS maximum channel information time. The output of the last channel pulse is linked back into the clock generator so that whenever a pulse is received it overrides the 10mS part of the cycle and triggers the 6mS generating circuit. The frame rate therefore varies from 60 frames per second for all the controls in the maximum position, to approximately 80 frames per second with the controls at minimum. With an average rate of 70 frames per second, the smoothness of servo response is greatly improved over the slower systems.

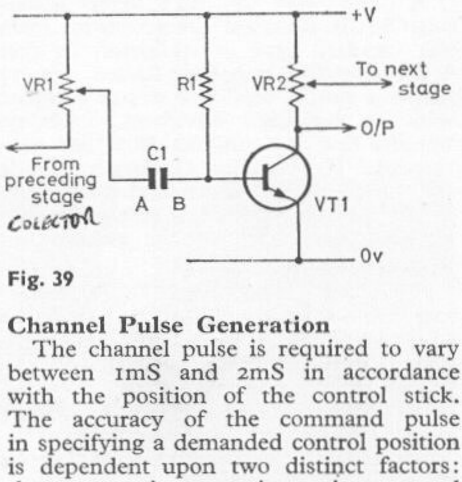


Fig. 39

Channel Pulse Generation

The channel pulse is required to vary between 1mS and 2mS in accordance with the position of the control stick. The accuracy of the command pulse in specifying a demanded control position is dependent upon two distinct factors: the accuracy in converting a given control stick position to a voltage and the accuracy of the circuits which convert the voltage to a pulse width. The circuit designer has very little control of the first case since the accuracy in this area is a function of the potentiometer linearity and the mechanical precision of the control stick assembly.

A typical channel stage is shown in Fig. 39. When the input voltage falls the transi-

tion that occurs on plate A of capacitor C1 is a proportion of the total voltage transition, dependent upon the position of the wiper. Plate B falls by the same amount causing the transistor to stop conducting. Capacitor C1 charges up through R1 and the transistor starts conducting again when its turn on threshold is reached. By varying the position of the potentiometer wiper, the magnitude of the transistion applied to the capacitor is varied, and hence the time for which the transistor is turned off varies. The waveform on the base of the transistor for two different control stick positions is shown graphically in Fig. 40. It should be

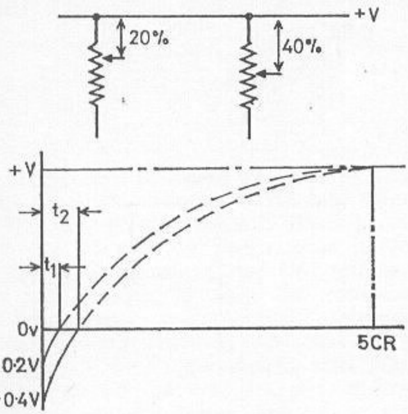


Fig. 40

remembered that the time constant is a function of C1 and R1, and that C1 charges towards +V irrespective of the magnitude of the transition applied to it. Thus for the two potentiometer positions shown, the transistor stops conducting for the period t_1 or t_2. A positive pulse of duration t_2 or t_2 is therefore produced at its collector.

It can be shown mathematically that for the circuit shown in Fig. 39 and for the following values,

C1 = 0.047microfarad
R1 = 125
+V = 10v

the potentiometer wiper must be at the following voltage below +V for the corresponding pulse widths.

1.0mS 1.8v
1.5mS 2.9v
2.0mS 4.1v

It should be noted that this represents an almost linear relationship between voltage and pulse width, and consequently for a linear resistance track the pulse width varies linearly with control stick position.

Assuming that the transistor driving the potentiometer saturates there will be approximately 10volts across the complete resistance track. To achieve a variation between 1.8volts and 4.1volts the wiper is required to operate over 23% of the track length. Since most potentiometers have a 270° track length, this represents 62° of control stick movement, with the neutral position 29° from the top end of the resistance track.

Since the input of the stage is connected to the output of the preceding channel circuit, the effect of a positive transition at the input must be considered. This occurs at the beginning of the channel pulse of the preceding stage. If no interaction between channels can be permitted, capacitor C1 must be able to charge fully during the minimum off time of the preceding stage. The charging time constant is formed by C1 and the resistance of VR1 between the wiper and the resistive voltage supply. Referring back to Fig. 22 it can be seen that a period of at least five time-constants is required in order to achieve a reasonable degree of stability, and so the time constant in this instance must not exceed one fifth of the minimum channel command pulse width i.e. 0.2mS. For a capacitor value of 0.047 microfarads, the maximum resistance must not exceed 4.25K.

It was shown above that the maximum control pulse width corresponds to the potentiometer wiper being 41% from the top end of the resistance track, and for the stability criterion to be satified this must not exceed 4.25K. A 10K potentiometer could be used but does not allow any margin for component tolerances. A 5K potentiometer is therefore employed giving a wide margin of safety.

A value of 125K was used for R1 in calculating the variation on the potentiometer wiper. If this is altered, both the neutral position and the range of pulse width variation for the given control stick movement are changed. R1 is there-

fore made variable to allow compensation for component tolerances, and can be used in conjunction with neutral adjustment of the control potentiometer for setting up each channel for identical performance. In practice R1 is a fixed resistor of 100K in series with an adjustable one of 50K maximum resistance.

The value of R1 imposes a minimum gain restriction on the transistor VT1. Sufficient collector current must be available to cause saturation. This is made up of the current through the potentiometer, approximately 2mA for a 10volt supply, and the current to the pulse shaping circuits and the following stage. With a transistor having a minimum gain of 100, the minimum collector current is 6.6mA since the base current cannot exceed 66 microamps in the worst case. This is about the limit that can be tolerated without the timing capacitor of the following stage causing degeneration of the trailing edge of the channel pulse. Modern silicon planar transistors are readily available having gain specifications in excess of this lower limit.

Accuracy

The two major factors that can cause inaccuracies in the width of the channel pulse are temperature and voltage variations.

Since the channel pulse width is determined by C1 and R1 (Fig. 39), any change in value with temperature will obviously affect the output. Disc ceramic and paper dielectric capacitors have poor temperature coefficients and so a moulded polyester or plastic film type is used for C1. For the resistor a high stability type is used, but it is not necessary to go to precision types since the temperature variation for resistors is several orders better than for capacitors.

Transistor gain tends to vary with temperature, so the minimum gain requirement should be applied at the worst case temperature. The simplest way of ensuring this is to allow a large safety margin on the transistor gain.

Voltage variation presents a different problem. If the voltage remaining across the transistor when saturated is ignored, then the pulse width is a function of the wiper position only. However if the saturation voltage is taken into consideration, then since this is reasonably constant, the voltage across the potentiometer does not vary in direct proportion with the supply voltage. The problem can be minimised by using low saturation voltage transistors. If a greater accuracy is required it is necessary to use zener diode stabilisation of the supply voltage with a resultant increase in current consumption.

A further consideration is the corresponding timing circuits in the servo amplifier. If the transmitter is voltage stabilised and the servo is not, then errors will be introduced as the servo supply voltage falls. Now the circuits employed in the servo, have similar characteristics to the transmitter encoding stages, consequently the timings change in the same sense as the supply voltage drops. A system without stabilisation therefore achieves a sufficient order of accuracy provided that the transmitter and servo supplies are used after having been charged together. A typical system accuracy of 3% can be achieved over the full voltage range of a nickel-cadmium cell type of supply, and the majority of the variation is accounted for by the high voltage which exists immediately after recharging. This high value quickly falls off due to the characteristics of the cell, and so the errors introduced only occur for a short initial period during the cell discharge cycle. A typical characteristic of a nickel-cadmium cell is shown in Fig. 41.

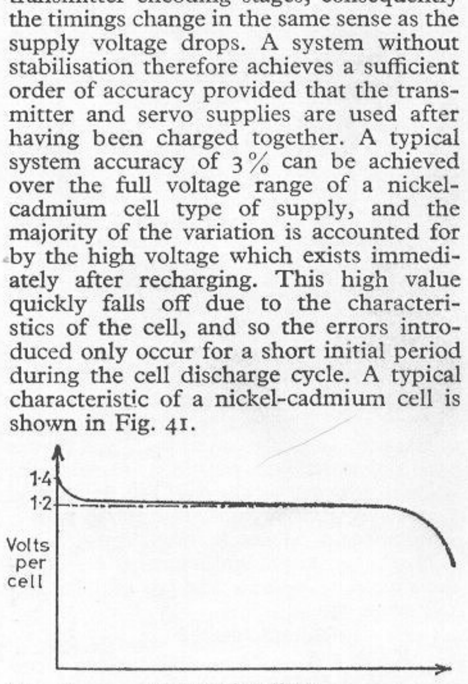


Fig. 41

If this type of operation is not acceptable, zener diode stabilisation must be employed in both the transmitter and servos. In the latter case the extra current consumption is hardly justified, particularly for a system intended for aircraft use.

The use of integrated circuits for transmitter encoders has been somewhat limited until the introduction of CMOS logic gave a circuit element of low power consumption and capable of operating on a wide range of supply voltages. Whilst it is possible to replace conventional discrete circuits by a chain of integrated circuit monostable pulse generators, this offers little advantage in terms of cost and component saving and so has not found much commercial application.

The more ingenious circuit designers tend to use a single pulse generator circuit and a counting system that switches in the control potentiometer for each channel in sequence. Such a system requires the selection of a good quality potentiometer if variations between channels due to track wear and manufacturing differences are to be minimised.

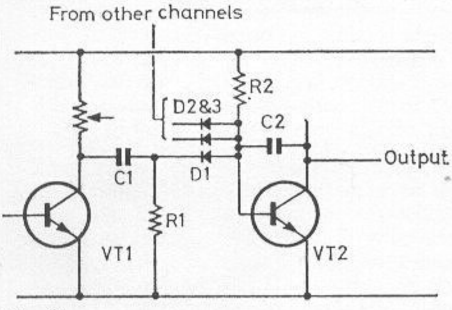


Fig. 42

The Modulation Pulse

There are many variations on the method of producing modulating pulses from the channel command pulses, and two of these will be discussed more fully.

Fig. 42 shows a simple method such has been employed from the earliest digital

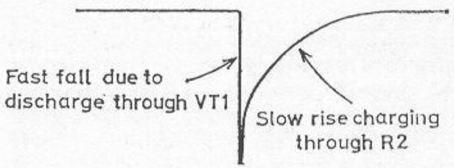


Fig. 43

systems. VT1 is a channel pulse stage, and VT2 is the transistor which is used to modulate the R.F. circuit. When a negative transistion occurs on the collector of VT1, capacitor C1 is discharged with the result that transistor VT2 is turned off via D1. The capacitor C1 then charges up through D1 and R2 allowing VT2 to turn on again. The time constant is so chosen that the output pulse width at the collector of VT2 meets the requirements for bandwidth. Diodes D1, D2, D3 etc. and R2 form a diode OR gate to allow the modulating pulses from further channels to be multiplexed.

The shape of the modulating pulse produced by the differentiator circuit is shown in Fig. 43, and unless care is taken in the modulator, this can cause "splatter" in the transmitter output. Capacitor C2 in Fig. 42 slows the turn on and turn off of VT2 as a measure towards reducing the bandwidth.

The second method of generating modulating pulses incorporates refinements to allow accurate control of the pulse shape. This is illustrated in Fig. 44.

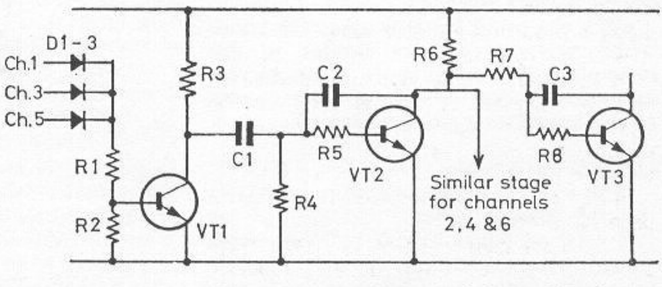


Fig. 44

Alternate positive going channel pulses are applied to the OR gate D1 D2 D3 R1 R2, and are shaped and inverted by VT1. C1 and R4 form a differentiator and operate on VT2 in the same manner as des-

cribed in the earlier circuit except that here they are used to turn the transistor on. C2 slows the rise and fall times and R5 is included to swamp variations in the input resistance of the transistor. A similar circuit is used to handle the pulses from the other channels, and the differentiated pulses occur at the time instants not defined by first circuit. Hence if the second circuit is also connected to R6, the full information will be available at this point.

VT3 is in a state of conduction except when a modulating pulse occurs to turn it off. Since it has a turn on threshold, when the signal at R6 falls below a certain level conduction stops, and restarts again when the threshold is reached. The shape of the pulse on R6, and the conduction threshold, therefore define the pulse width. The C3, R8 combination is included to slow the edges since VT3 would otherwise tend to speed up the shaped edges at R6. The particular advantage of this circuit is that the pulse shaping circuits are not influenced by the load applied to the modulating transistor to the same degree compared with the earlier example. The only factor which is influenced by the load circuit is the pulse width since the threshold level will be altered. This can be readily compensated by choosing the value of R7 to give the appropriate amount of base current when the differentiated pulses are at the required voltage level.

With the trend towards narrower bandwidth systems, circuits similar to this example will become more commonplace as manufacturers aim for greater control of the modulating pulse waveform.

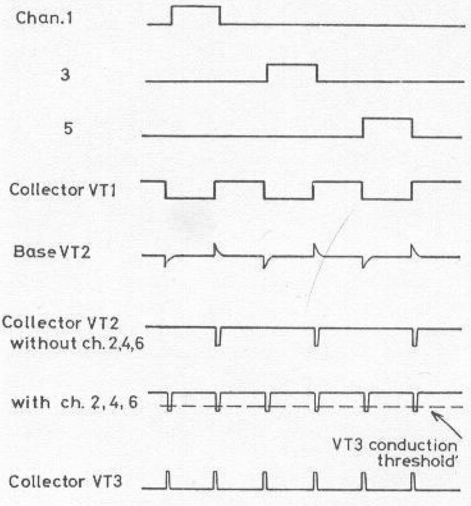


Fig. 44a

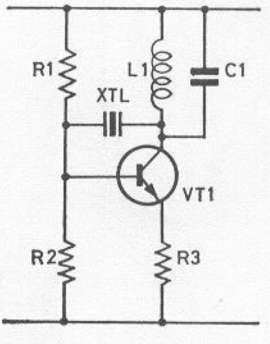


Fig. 45

Crystal Oscillator

The 27MHz carrier signal is invariably derived from a crystal oscillator circuit and then amplified by one or more stages before being driven into the aerial circuit. The modulating signal may be impressed upon the carrier at any one of the stages in the system, and the various considerations on how this is achieved will be discussed later.

The master oscillator usually takes the form shown in Fig. 45. The crystal provides a low impedance circuit only to signals at its resonant frequency. Thus if a transient of the correct frequency occurs at the collector of the transistor, it provides positive feedback and causes the cycle of oscillation to be completed. The resonant circuit formed by L1 and C1 serves both to start the circuit oscillating and to provide a means of taking an output from the stage. If its resonant frequency is exactly the same as that of the crystal the voltage developed across it will provide the maximum feedback through the crystal. If the two frequencies differ slightly the tuned

circuit will be dragged to the frequency of the crystal since the transistor amplifies this frequency to a much higher power level than the L-C circuit can provide. The signal developed in the tuned circuit will therefore be at the crystal frequency but will suffer slight distortion. This is not usually of great significance since the following stages incorporate tuned circuits which will select only the fundamental frequency component.

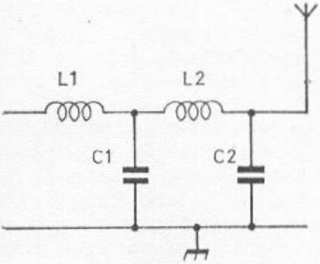


Fig. 46

Power Amplifiers

The power amplifier stage which provides the final output includes the aerial in its resonant circuit. A straight conductor has a certain inductance and capacitance dependent upon its length and hence has a resonant frequency, so can serve as an aerial. Unfortunately at 27MHz, the length required is far too long to be used in practice, so a short length is used in conjunction with additional inductance and capacitance. The input impedance of a tuned circuit is a function of both the inductance and capacitance and so the values chosen when designing a circuit, have to take the output impedance of the driving stage into consideration for maximum power transfer to be achieved.

Various configurations are employed for achieving a resonant circuit. Sometimes a coil is inserted in the aerial itself (centre loading) to give increased inductance, and capacitance added at the base. Alternatively the inductance is added at the base (base loading) together with the capacitance. A typical base loading network is shown in Fig. 46. By making both C1 and C2 variable both the input impedance and resonant frequency can be readily adjusted.

From a practical consideration in modelling applications, the base loaded aerial is to be preferred since it is less prone to accidental damage, and can be interchanged for another of equal length. An error in length of several centimetres will represent only a small change in its characteristics and the output will not be greatly affected. With a centre loaded aerial a replacement is susceptable to tolerancing on the coil, and since the circuitry at the base is simpler, the length becomes a more critical factor if R.F. output is not to be affected.

The aerial circuit is driven by a transistor amplifier which in one form can be as shown in Fig. 47. L1 represents transformer coupling from the preceding stage, and is preferred to other methods since impedance matching can be readily

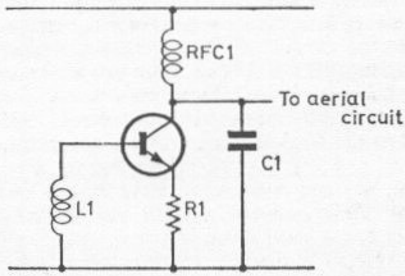


Fig. 47

achieved by adjusting the number of turns.

R.F.C.1 allows d.c. current to reach the stage whilst blocking the R.F. signal from the supply line. R1 is a current limiting resistor to protect the transistor from damage due to excessive current. Such a condition might otherwise occur during the tuning proceedure, or when the aerial is removed so causing mismatching of impedance. The capacitor C1 is not strictly necessary since it partially contributes to the aerial circuit. It is included to swamp

the output capacitance of the transistor, a somewhat variable factor, and the aerial circuit is designed taking this into account.

Another form of output stage uses two transistors with their outputs commoned (Fig. 48). This is really only a variation on the preceding circuit to allow lower power rating transistors to be employed. Separate emitter resistors are required in order to produce current sharing between the two transistors and hence avoids one transistor carrying more than half the signal power. The use of this circuit is purely a factor of production costs, since high frequency low power handling transistors are considerably cheaper than high power rated types.

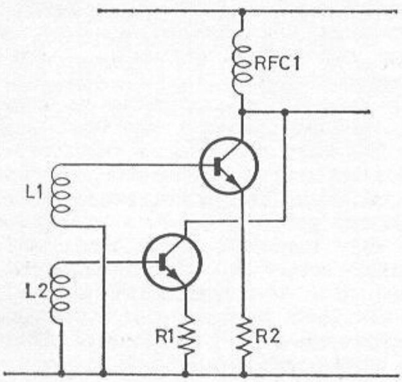


Fig. 48

The third type of circuit that has been employed uses two transistors in "push-pull" configuration (Fig. 49). By correctly phasing the coupling coils on the input, one transistor handles the positive half of the 27MHz oscillation cycle, and the other transistor handles the negative part. The peak to peak output voltage swing can approach twice the d.c. supply voltage without distortion being introduced due to voltage clipping. However the set up proceedure for a circuit of this type is far more critical since both halves of the oscillation cycle have to be set up to be identical and also the transition from one to the other must exhibit no discontinuity. Because of this disadvantage, the push-pull circuit is not normally found in digital systems.

Inputs to Power Amplifiers

A power amplifier stage will have a certain power gain, and if a certain power level is required at the output, the gain determines the input power required. The power level that can be obtained from a crystal oscillator circuit of the type normally employed, is adequate to drive a medium to high gain power amplifier and provide sufficient R.F. output. However to achieve the required order of gain with

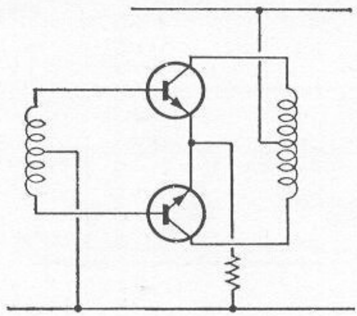


Fig. 49

economically priced transistors, means that the power amplifier design is working towards the limits and could be subject to instability. It is far better practice to include a driver stage between the crystal oscillator and power amplifier, particularly if this includes a tuned circuit and is transformer coupled to allow optimised impedance matching. A further advantage will also become evident in the next section.

Modulating the Carrier Signal

The control information is in the form of pulses and it is required that these should turn the R.F. carrier on and off. This can be done either at the crystal oscillator, or at the power amplifier, or at any intermediate stages that might be included.

If the crystal oscillator is switched by the modulating transistor, there is a tendency for it not to turn on cleanly due to the change in impedance caused by the characteristics of the oscillator transistor altering with collector current. This results in transmitter splatter, i.e. the introduction of incorrect frequencies into the

output. Since this is somewhat dependent upon individual transistor characteristics, this can cause difficulties in setting up an R.F. circuit.

An alternative approach is to switch the power amplifier. This is a perfectly satisfactory approach except for the modulating transistor having to be capable of switching the maximum current required by the power amplifier. The danger in this case is that if the switching is not complete poor modulation results and can cause erroneous reception conditions in the receiver with changing range.

The most satisfactory modulation method is to switch a driver stage before the power amplifier. This avoids the splatter problem associated with the crystal oscillator, and only requires currents in the order of a few milliamperes to be switched. The tuned circuit in the driver stage will also provide a degree of filtering to reject any unwanted frequencies that might have been introduced by the modulation pulse. With the trend towards narrower bandwidth R/C systems, circuits of this type should become more commonplace.

Microtrol-5 Digital Transmitter

Having considered the various aspects of circuit design, the Microtrol-5 transmitter will be discussed to give an example of how the features may be combined into an overall unit.

Fig. 50 shows the complete circuit and the operation will be followed through stage by stage.

VT9 and VT10 form the basic encoder clock with the time constant formed by VR2, R23 and C25 providing the sychronisation pulse width. In normal operation, the last pulse in the command sequence, produced at VT15 collector, is fed back via R24 to provide a negative transition on C25 to initiate this time period. If no transition is produced via this path e.g. when first switching on, the longer time constant R25, C27 causes the cycle to be started. Diodes D1 and D2 protect the transistor emitter—base junctions from breakdown due to the large voltage transitions that occur. Typically the transistors can only tolerate the base being approximately 6volts more negative than the emitter, and the transitions are in the order of 9 volts. VR8 is the first control potentiometer, and the first channel pulse is timed by C29, VR3 and R26. Since the potentiometer wiper never passes below the halfway position for a 1 to 2 millisecond control pulse, the voltage transition will not be sufficiently large to cause damage to VT11. VR3 is a preset control for setting up the stage. VR9 is the control potentiometer of the next stage. The channel stages are repeated up to VT15 where the collector load becomes a fixed resistor since no further controls are required.

The small value capacitor (0.001 microfarad) from the base and collector of each transistor are included to bypass any R.F. signal that might be picked up particularly by the leads to the control potentiometers.

VT1, 2, 3 and 4 form the modulator stage, one side being driven via D3, D5 and D7, and the alternate pulses from D4, D6, D8 driving the other side. It should be noted that if an even number of channels is required, the variable frame rate system cannot be used with the modulator in this form, since the same side of the modulator would need to handle two successive pulses. However, with an odd number of channels as in this five channel system, there is no problem.

The modulator operates exactly as detailed earlier in this chapter, with the differentiator being formed by C2 and the circuits to 0v via R5 in series with the emitter junction, and R4. C3 is the roll off capacitor to smooth the transitions at the collector of VT2. An OR function between the two groups of channel pulses is achieved by using R6 as a common collector load to VT2 and VT3.

The crystal oscillator stage, VT5, L1, C8 etc., is decoupled from the supply lines by RFC1 and C7. This is particularly necessary since if any of the R.F. stages were to pick up extraneous signals from each other they would have an incorrect phase relationship with the correct signal, and produce distortion in the output. Resistors R12, 13, 14 provide d.c. bias conditions for VT5.

A tuned R.F. driver stage, VT6, is modulated by VT7. Decoupling of the

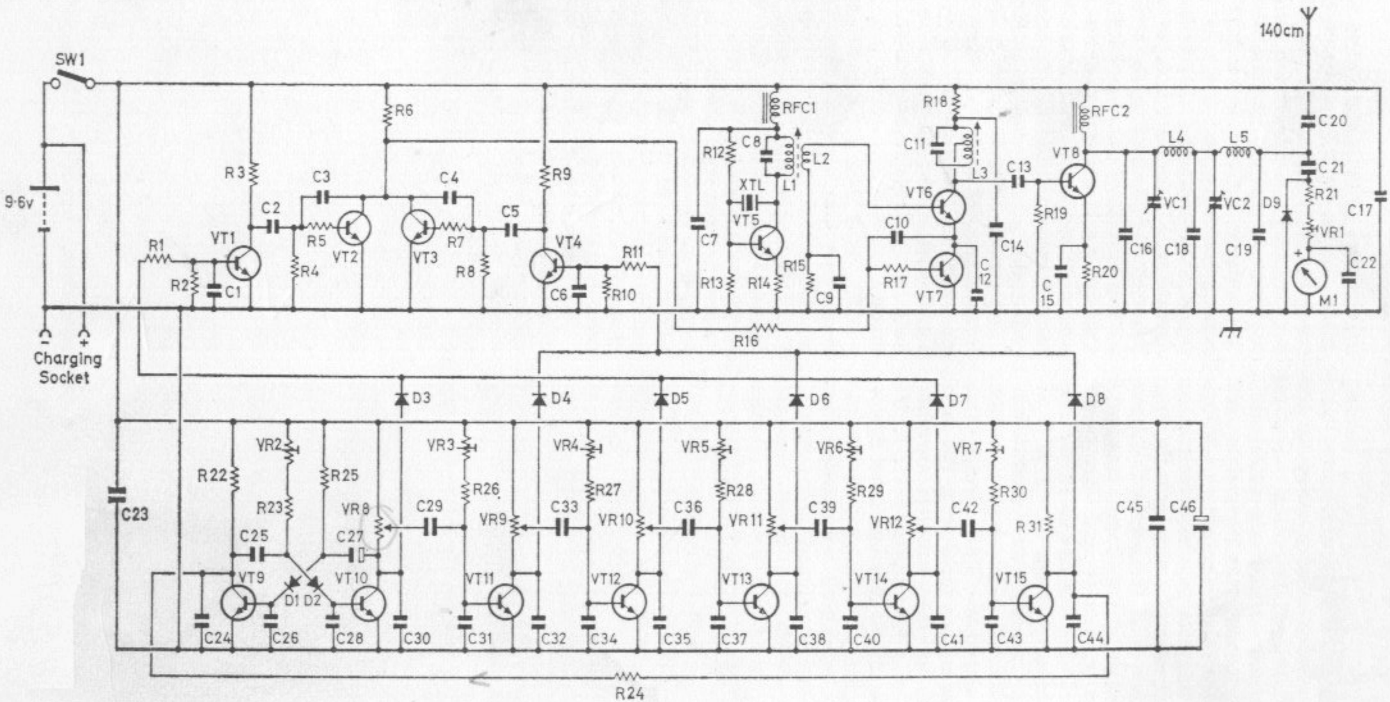


Fig. 50 MICROTROL-5 DIGITAL TRANSMITTER

stage from the supply is by means of C14 and R18, the latter also serving to limit the d.c. current in the circuit. L2 provides an impedance input for the local oscillator output, and C9 produces an a.c. bypass path to R15 so that for the R.F. signal, the lower end of L2 is effectively at the 0V reference. C12 performs a similar function to decouple the modulator switch VT7.

The tuned circuit C11, L3 is impedance matched into the power amplifier by the tapped coil. This allows capacitive coupling to be used for driving the final stage. R20 is a d.c. current limiting resistor for VT8 and R.F. decoupling is provided by C15. RFC2 allows d.c. current to flow in the circuit but blocks the R.F. signal form the supply line. The aerial tuned circuit comprises C16, VC1, L4, C18. VC2, L5, C19 and the aerial. Low value adjustable capacitors are used to provide a fine adjustment to the paralleled fixed capacitors. C20 is included solely to provide a block to d.c. current; since the aerial might accidentally be shorted to the 0V reference which is connected to the transmitter case.

Meter M1 indicates the R.F. signal voltage applied to the base of the aerial. C21 couples a small proportion of this voltage into the circuit, and it is half wave rectified by D9. VR1 and R21 determine the calibration and C22 provides a filtering action.

Low frequency decoupling of the supply lines is provided by the electrolytic capacitor C46. Disc ceramic capacitors C17, C23 and C45 produce the same function at high frequencies and are used at strategic points in the circuit dependent upon the printed circuit layout.

BLANK

CHAPTER 7

DIGITAL RECEIVERS

POPULAR demand dictates that modern proportional equipment shall be capable of operation alongside other outfits all operating within a narrow frequency band. In the U.K. the licensing authorities allow operation within the range 26.96MHz to 27.28MHz. This is similar to many other countries although some have a tighter restriction in that specific frequencies are allocated. As a result the frequencies in the following table have become a world wide standard.

MHz	Colour Code
26.995	Brown
27.045	Red
27.095	Orange
27.145	Yellow
27.195	Green
27.245	Blue

Increasingly, however, outfits are operated on the so-called "split" frequencies—that is a frequency approximately midway between the agreed six. These are not specified here because, currently, there is no precise agreement and they vary from an exact 25MHz split, to a 20-30 MHz split, designed to cater for future 10 MHz spacing. The split frequency identification is normally by a flag carrying the two colours from either side—e.g. a set on 27.020 would have a red/orange flag. It should be noted that the " blue " frequency is just as liable to be on 27.255 MHz, as on the " correct " 27.245MHz.

With frequencies in such close proximity, superhetodyne receivers are essential if simultaneous operation is to be achieved.

In earlier chapters the implications of bandwidth have been considered, and if reliable operation is to be achieved, the selectivity of the receiver is a most important factor. With 25KHz channel spacing, an extremely narrow bandwidth is required if the receiver is to accept only the correct signal under all conditions of varying signal strength. It is in this aspect that receivers for proportional equipment differ greatly from those produced for any other application. In broadcasting, if an adjacent channel breaks through at a low level, it causes annoyance but is not catastrophic. With reed type multi channel R/C equipment, the filter bank provided additional selectivity. It is not surprising therefore to find that improvements in the reliability of proportional are closely related to receiver development, since selectivity must come from the R.F. circuits alone.

The Superheterodyne Principle

The principle of superheterodyning is that two slightly differing frequencies are mixed together in such a way that the result contains components at frequencies related to those of the inputs.

This form of mixing occurs when the input signals are passed simultaneously through a stage which has non-linear amplifying characteristics. The most prominent components in the output occur at frequencies representing the sum and difference of the inputs. Thus for an input of 26.995MHz mixed with a signal of 26.540MHz, strong components are produced at 53.535MHz and 455KHz. (1000KHz=1MHz). Any modulation superimposed upon either of the input signals will also be present in the output.

Another factor that must be considered is that the same difference frequency of 455KHz can also be produced by an incoming signal of 26.085MHz mixing with

the 26.540MHz local oscillator. Whilst in most areas in the U.K. no difficulties exist due to frequency band allocations, in the U.S.A. there are some areas where transmissions do occur on frequencies that could cause difficulties. For this reason some American equipment uses receiver local oscillator frequencies higher than the incoming signal, whereas in the U.K. the lower of the two frequencies is almost universally used for the local oscillator.

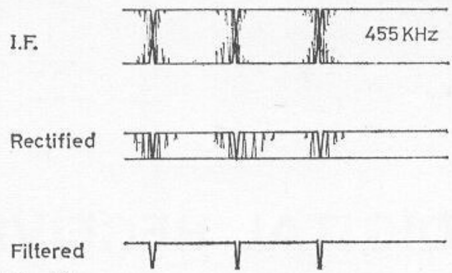


Fig. 52

Superheterodyne Receivers

The heart of any superhet. receiver is the I.F. strip comprising a mixer and the I.F. amplifiers. The stages are designed to be selective to one of the output frequencies, usually the difference frequency, resulting in a high output level signal at this frequency only. Fig. 51 illustrates this along with the associated areas of the system.

The R.F. detector incorporates a tuned circuit which includes the aerial. The 27MHz output from this stage is coupled into the mixer, usually by transformer coupling from the aerial coil. The local oscillator provides the second input frequency and is usually crystal controlled, an almost essential requirement for stability to be maintained.

The transistor in the mixer stage is biased to operate in a non linear region of its characteristics. The output of the stage is tuned to the difference frequency by means of the tuned primary of the first I.F.T. (Intermediate Frequency Transformer). Thus the other frequencies produced by the mixing process are rejected. It should be noted that by tuning to the difference frequency, the lowest intermediate frequency has been selected.

The two I.F. amplifier stages have the tuned primary windings of the second and third I.F. transformers as the collector loads of their transistors. Hence the signal that is available at the input to the detector stage has gone through several stages of selection and the total bandwidth is very small. A bandwidth of 5KHz for a 27MHz signal represents a degree of selection of 1 part in 5400.

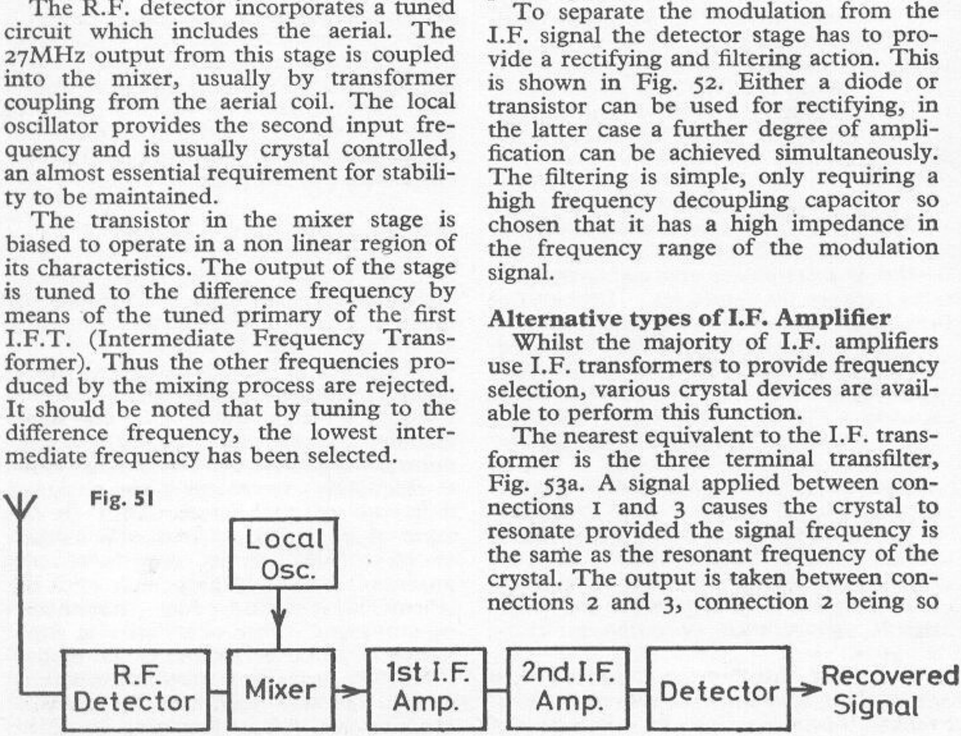
To separate the modulation from the I.F. signal the detector stage has to provide a rectifying and filtering action. This is shown in Fig. 52. Either a diode or transistor can be used for rectifying, in the latter case a further degree of amplification can be achieved simultaneously. The filtering is simple, only requiring a high frequency decoupling capacitor so chosen that it has a high impedance in the frequency range of the modulation signal.

Alternative types of I.F. Amplifier

Whilst the majority of I.F. amplifiers use I.F. transformers to provide frequency selection, various crystal devices are available to perform this function.

The nearest equivalent to the I.F. transformer is the three terminal transfilter, Fig. 53a. A signal applied between connections 1 and 3 causes the crystal to resonate provided the signal frequency is the same as the resonant frequency of the crystal. The output is taken between connections 2 and 3, connection 2 being so

Fig. 51



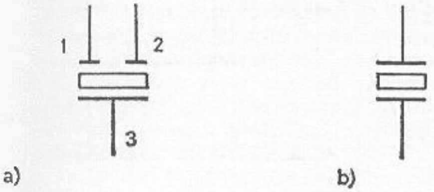


Fig. 53

positioned on the crystal slice that it provides a low source impedance to the following stage. A second type of transfilter has two connections, Fig. 53b. At its resonant frequency it provides a low impedance and so can be used as a decoupling element. Fig. 54 shows part of an I.F. amplifier using both types of transfilter.

There are several problems associated with using transfilters in R/C receivers. Compared with an I.F. can, the three connections transfilter is larger and fragile and more prone to vibration. Apart from this, since they are not tuneable they present a tolerancing problem from a manufacturing viewpoint. (For domestic radio applications the local oscillator is tuneable to provide a means of station selection, and so avoids this problem). The two connection transfilter is sometimes used in place of emitter decoupling capacitors in all types of I.F. amplifiers and does offer slightly increased selectivity. This has to be weighed against the cost and reliability factors, and with the high selectivity of modern miniature I.F. transformers any advantage is minimal.

An alternative approach is the use of filter blocks. These are fairly complex crystal filter networks made to tight tolerances and closely defined characteristics. It is possible to achieve a very sharp frequency cut off either side of the specified bandwidth but the manufacturing cost is high if the specifications are to be maintained. If the frequency is not tightly controlled low sensitivity will result due to the difference frequency generated in the receiver being outside the filter block pass band. Whilst the filter block replaces the I.F. transformers functionally, it is still necessary to employ tuneable transformers to achieve impedance matching between the filter block and the external circuits. Hence whilst offering advantages in terms of electrical performance, mechanically a filter block is no more robust than a set of conventional I.F. transformers

Automatic Gain Control

Any form of amplifying circuit can only operate correctly over a specific range of signal voltage levels. With a radio control receiver no malfunction can be tolerated over the entire range of operation, including the transmitter and receiver aerials in close proximity. To accommodate this large range of signal levels a receiver circuit is arranged so that as the signal level increases, the gain is reduced. This is termed automatic gain control (A.G.C.). The great majority of broadcast receivers incorporate this feature to minimise changes in volume due to signal fading. Occasionally on long distance stations the signal can fade due to atmospheric conditions, below a level where the A.G.C.

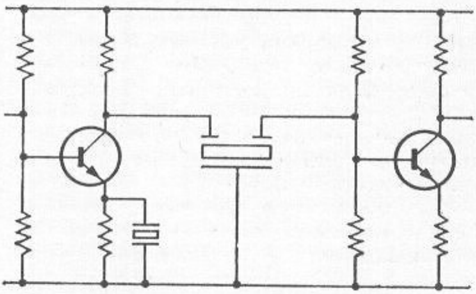


Fig. 54

can compensate. This is evidenced by a reduction in the output volume.

With a radio control receiver the same problems exist but with the major difference that loss of output could be catastrophic. Additionally the range of signal levels that has to be accomodated is far greater than that for which any broadcast receiver is expected to compensate, and the response speed of the circuits must be sufficiently fast not to produce a momentary loss of signal, as the model passes in close proximity to the transmitter. The design of the A.G.C. circuit is therefore a critical factor in the performance of the receiver.

In chapter 5 when discussing amplifier

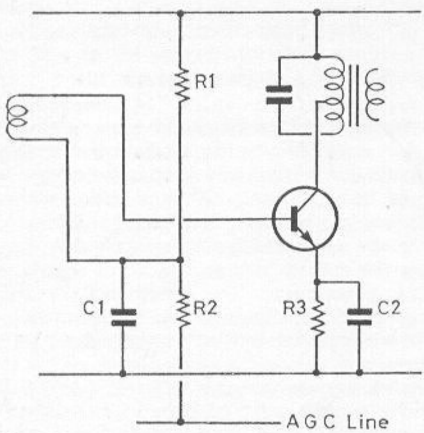


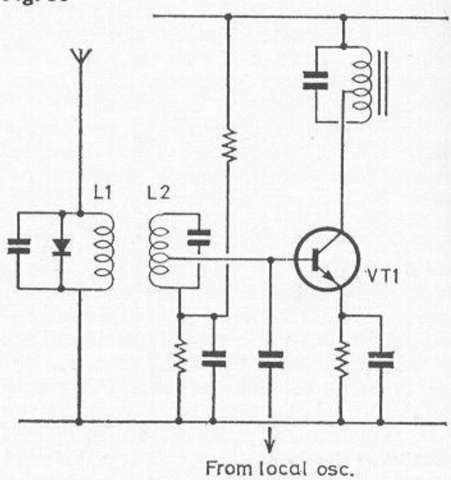
Fig. 55

circuits, it was shown that the gain of an amplifier stage was related to the biasing conditions. The gain can therefore be readily controlled by returning a base bias resistor to a voltage level that varies with the signal strength. A typical I.F. amplifier stage is shown in Fig. 55. As the AGC line moves towards the lower supply line, negative for an NPN transistor, the forward bias on the transistor decreases. This causes less current to flow into the collector circuit, so resulting in a reduction of the overall gain of the amplifier stage.

The voltage applied to the A.G.C. line has to be derived from a point in the circuit that exhibits some form of change with varying signal levels. The detector stage (Fig. 51) has a rectifying action causing a d.c. voltage to be established that is dependent upon the signal strength. The recovered modulation signal is superimposed upon this d.c. voltage and must be filtered out before it can be used to control the I.F. amplifier gain. If a filter is not incorporated, the high level of the modulation signal would tend to swing the bias voltage outside of the range that the amplifier can handle. In Fig. 55, C1 performs the filtering action. The time constant formed by C1 and R2 may prevent the receiver operating correctly under conditions of rapidly changing signal strength, e.g. when a fast moving model passes close to the transmitter. Some early receiver designs were prone to "glitch" under these conditions. Apart from this problem, sufficient control of receiver gain can be achieved by applying A.G.C. to one I.F. amplifier only.

A considerable improvement can be effected by applying a smaller amount of A.G.C. bias change, but to more than one stage, so that the overall receiver gain varies over the same range. Since the proportion of signal coupled back is recuded, the tendency for the modulation to cause saturation of the amplifier is minimised and the filter time constant can be made shorter. The Microtrol receiver discussed later in this chapter employs this A.G.C. technique, applying a bias to the mixer stage as well as the two I.F. amplifiers. A feature sometimes used to prevent excessive signal levels entering the receiver, is to provide a diode across the aerial tuning coil, (Fig. 56). Signal levels in excess of the forward conduction voltage (about 0.7 volt for a silicon diode) are clipped as the diode starts to conduct, so do not cause overloading of the following circuits. This technique is usually only employed with a double tuned front end,

Fig. 56



since the carrier signal distortion introduced by clipping is removed by the tuned coupling between the coils.

Noise Rejection

Whilst the R.F. circuits and superheterodyne provide rejection of unwanted R.F. signals, system malfunction can occur due to locally introduced electrical noise. It is not unduly difficult to prevent spurious signals entering the amplifier stages and decoder, but due to the high receiver sensitivity they can enter into the receiver itself and so are present in the recovered modulation signal, (Fig. 51). Such sources of noise are servo motors, metal to metal linkages and noise introduced from within the transistors themselves, the latter being caused by electron movement within the semiconductor crystal structure.

Metal to metal noise enters the receiver via the aerial and its major frequency components will be below 27MHz. By using a double tuned R.F. front end (Fig. 56), noise from this source is virtually prevented from entering the system, since there is no d.c. path directly into the first transistor stage. A simpler system, eliminating L1 and coupling the aerial into L2 allows any low frequency noise picked up by the aerial to be passed to VT1.

Servo motor noise can enter either via the aerial or via the battery supply lines. Interference picked up by the aerial can be overcome by fitting high frequency suppresion across the servo motor brushes. The receiver can be made insensitive to low frequency transients on the supply lines by capacitor-resistor decoupling of the lines at the receiver.

The net result of these noise sources is that our recovered signal has unwanted noise superimposed on it as shown in Fig. 57. Provided the noise spikes do not approach the same amplitude as the minimum signal level, separation can be

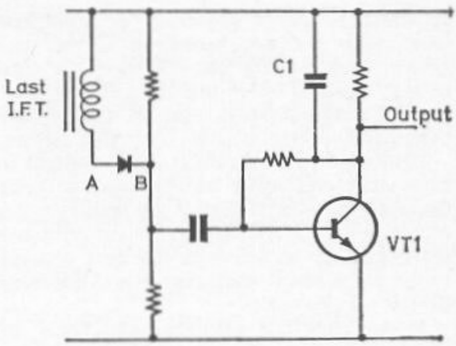


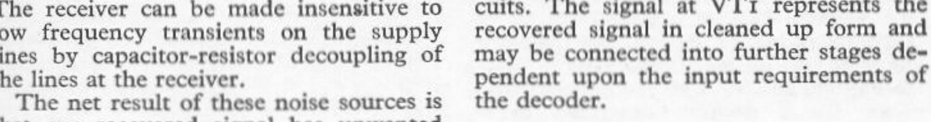
Fig. 58

achieved by incorporating a threshold detecting circuit. There are several methods of producing such a circuit but the most satisfactory method is a stage which does not include a transistor thereby eliminating any variation in performance due to gain tolerances. Fig. 58 shows such a method in which a bias voltage is applied to the detector diode. Before any output can be produced the signal at point A has to exceed the bias level at B, plus the forward conduction voltage of the diode (about 0.3 volts for a germanium signal diode). Low level noise present at A does not therefore appear at B. The output may then be capacitively coupled into a simple transistor stage as shown. C1 provides a decoupling action to filter out any high frequency noise that may have been passed through the circuits. The signal at VT1 represents the recovered signal in cleaned up form and may be connected into further stages dependent upon the input requirements of the decoder.

Design Considerations

Unlike transmitters, it is not easy to highlight the features which give one receiver design a superior performance to

Fig. 57



another. What is required, is sufficient bandwidth for the incoming signal, together with a sharp cut off on either side of this band. This is mainly a function of the I.F. transformers and so the circuit designer has little control over this aspect.

Provision of adequate A.G.C. control is important and must be achieved without derating the selectivity of the receiver. To a certain extent trial and error experiments are necessary to arrive at the best control range for a given transmitter and receiver design combination.

Noise rejection features contribute to eliminating "glitches", and since these represent an unrepeatable type of fault, the optimum circuit in a particular receiver design has to be determined empirically.

Temperature stability is again mainly a function of the I.F. cans, but the choice of transistors can be important. Electron noise in semiconductors is a function of temperature and if a poor choice of transistor is made, the variation in noise levels can upset the operation of even the best of noise rejection circuit design.

With such a large proportion of a receiver circuit not being the subject of mathematical design, it is not surprising to find that the majority of commercial receivers incorporate all the major features that over a period of time have proved to contribute to reliable performance.

Microtrol Receiver

As an example of the method by which the various features outlined can be incorporated into a circuit, the Microtrol receiver will be discussed. This is shown in Fig. 59, and includes the sychronisation logic for a decoder described in the following chapter. The diagram of the decoder integrated circuit is shown in Fig. 74 and its operation is discussed in detail on page 93. Constructional notes appear in Appendix II.

The flex aerial is connected to the primary of a double tuned R.F. circuit, tuning of this being by a fixed capacitor C2, and a variable inductance L1. The diode D1 limits the signal at resonance under strong reception conditions to eliminate any swamping tendency. L2 and C5 form the resonant secondary circuit, and inductive coupling to the primary is effected by positioning the coils so that their axes are parallel and in close proximity. Impedance matching into the mixer transistor VT2, is achieved by a tapping point on L2. VT1 and associated components form the local oscillator circuit, and is crystal controlled to a frequency 455KHz below the incoming signal. The inductance of RFC1, and C6 represent the resonant components in this circuit. C7 couples the local oscillator frequency into the mixer stage.

Mixing of the frequencies is performed by VT2 which is biased to operate as a non linear amplifier. The output of this stage feeds the primary of the first I.F. transformer, this being tunable to the difference frequency by the internal capacitor and an adjustable ferrite "core". The output from the untuned secondary, drives the first I.F. amplifier VT3, the gain of this stage being deliberately limited by the omission of a decoupling capacitor across its emitter resistor.

I.F.T.2 couples the signal into the second I.F amplifier VT4, which in turn feed I.F.T.3. The signal at the collector of VT3 is the maximum I.F. signal available within the receiver so is used to generate an A.G.C. voltage. C12 couples the signal to D2 which performs half wave rectification, the mean value of the remaining half cycle being proportional to the I.F. signal level. R1 and R11 establish the variable bias voltage and C1 provides a smoothing action. A.G.C. is applied to both I.F. amplifiers and the mixer by means of the base resistors R2, R7 and R9. A return path to reference ground, at signal frequencies, is provided in the base circuits by capacitors C4, C9 and C10.

Recovery of the modulation signal is performed by the detector diode D3 which has a bias voltage applied to it for noise rejection purposes. C13 filters off the I.F. frequency from the recovered signal.

VT5 is the first amplifier stage and is biased in such a way that feedback is applied to the transition appearing at its base via the coupling capacitor C16. This slows down the rise and fall times of the pulses appearing at its collector in order to minimise any overshoot that might occur. C17 serves to decouple any fast noise

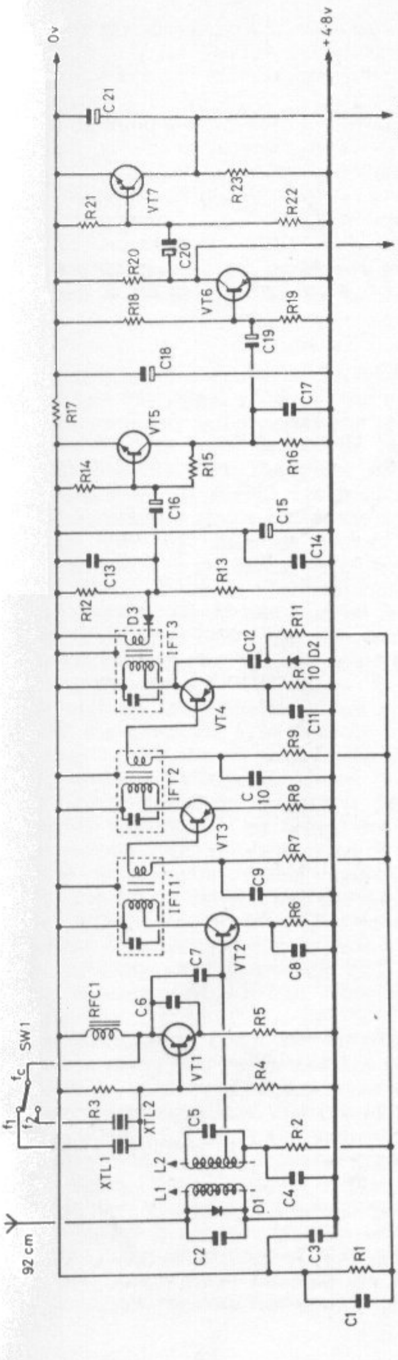


Fig. 59 MICROTROL RECEIVER

spikes that may have been passed through the receiver.

The entire receiver, including this amplifier stage, is decoupled from the supply lines by means of R17, C15, C14 and C3. This prevents any noise on the supply lines being picked up by the receiver. C3 and C14 are disc ceramic capacitors which decouple at signal frequencies since the electrolytic capacitor C15 is not efficient in the higher frequency range.

The output from VT5 is capacitive coupled into the succeeding stages by C19. VT6 and VT7 are used to generate signals required to operate the decoder, and their functions will be covered in the following chapter. It may be noted that VT6 operates as a saturating amplifier, providing a squared up version of the recovered signal pulse.

CHAPTER 8

DIGITAL DECODERS

THE function of a decoder is to take as its input the signal pulses at the output of the receiver, and separate the command information onto individual lines for each channel. With the pulse width modulation used in digital systems, the channel output is a single pulse, the start and finish times of which are defined by short pulses in the incoming pulse train. The waveforms shown in Fig. 30 illustrate how the encoded signal was derived from the individual channels. The decoder is required to perform the reverse process in separating the channel information.

It is perhaps the decoder which gives the name digital to this type of radio control system, since the circuits found in a decoder are purely digital in nature.

The basic principle employed in a decoder of this type is to count the incoming pulses. Thus the first pulse of the incoming information frame causes a count of 1 to be established, and the second pulse establishes a count of 2. Channel one is therefore defined by the time period for which the count of 1 is present and the process is continued for the remaining channels.

A circuit that can latch with incoming pulses must have two stable states if it is not to return to its original condition due to internal considerations. Fig. 60 shows such a configuration using two transistors. If VT1 is conducting, its collector current causes a voltage drop across R1 so that the voltage at the collector is approximately 0v. The voltage at the base of VT2 cannot exceed the collector voltage of VT1 since VT2 base current flows through R1 and R2. VT2 cannot conduct under these conditions and so its collector voltage is approximately +V. This allows base current for VT1 to flow via R3 so maintaining VT1 in a state of conduction. This represents one of the circuits two stable states i.e. VT1 conducting, VT2 turned off.

Since the circuit is symmetrical, a second stable state is represented by the condition of VT2 conducting and VT1 turned off. The circuit can only be changed

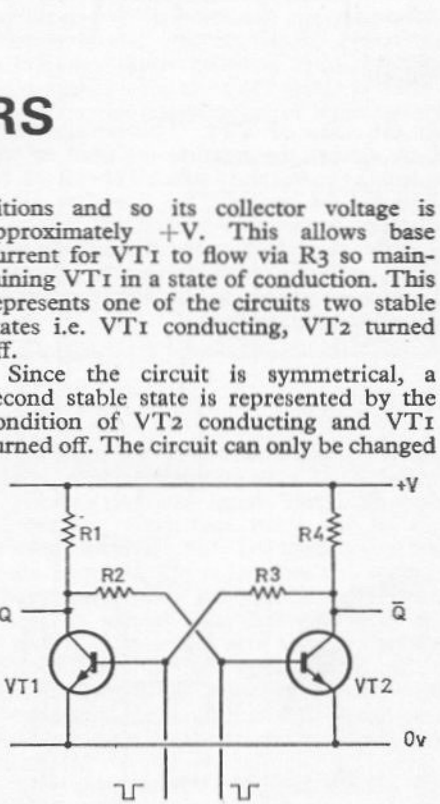


Fig. 60

from one state to the other by an external influence and this can conveniently take the form of a negative pulse applied to the base of the conducting transistor. A circuit of this form is termed a bistable and is one of the basic elements used in producing digital logic systems.

In order to make the bistable into a counter, the incoming pulses have to be steered alternatively between the bases of the two transistors. If VT1 is conducting the next negative pulse must be applied

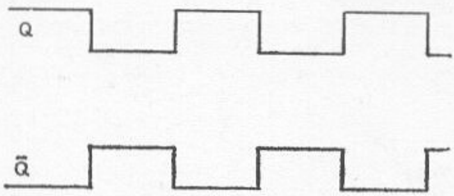


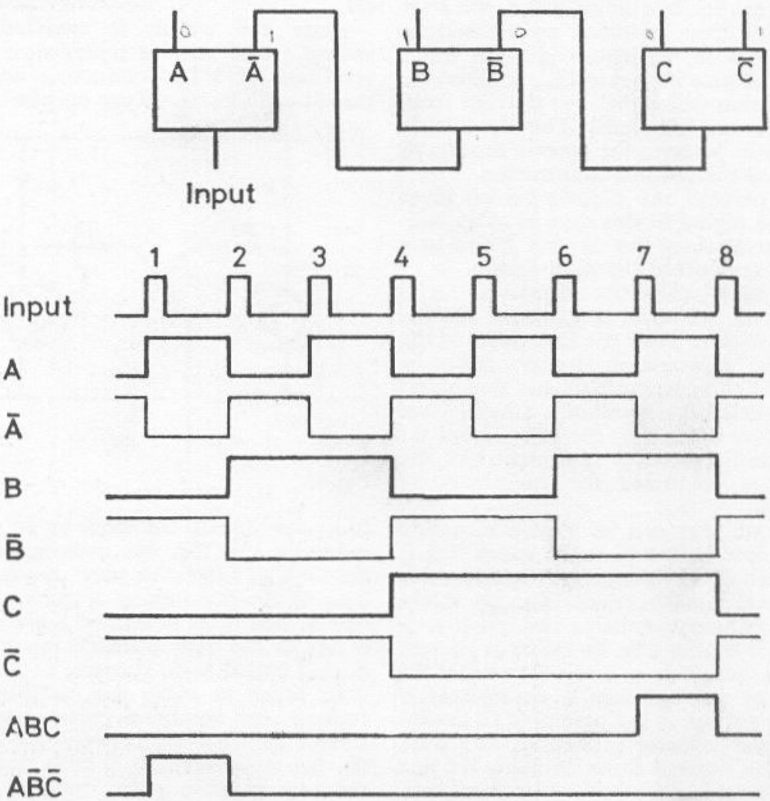
Fig. 61

to the base of VT1. This requirement fully defines the steering required on the incoming pulses and a suitable circuit can be included to produce the function. With discrete component assemblies this will usually be a diode-capacitor-resistor network, whereas with integrated circuits, transistor-resistor circuits are used within the encapsulation itself.

With integrated circuits, various gating circuits can be included on the inputs to produce a range of "flip-flops", to use normal terminology, having different input characteristics. When a single flip flop is interconnected to produce the counting function outlined above, it becomes one stage of a binary counter. The two possible outputs obtainable from the two collectors are complementary, i.e. $\bar{Q}$ is the inverse of Q as shown in Fig. 61. Thus a bar above a function name means that it is the inverse of the function.

If flip flops, having the property of changing state on the positive edge of the incoming pulse, are connected as shown in Fig. 62, this forms a binary counter. The waveforms at the outputs of the collectors are also shown in the diagram. The table shown in Fig. 63 is called a

Fig. 62



A(1)	B(2)	C(4)	Decimal	AND Gate
0	0	0	0	$\bar{A}\bar{B}\bar{C}$
1	0	0	1	$A\bar{B}\bar{C}$
0	1	0	2	$\bar{A}B\bar{C}$
1	1	0	3	$AB\bar{C}$
0	0	1	4	$\bar{A}\bar{B}C$
1	0	1	5	$A\bar{B}C$
0	1	1	6	$\bar{A}BC$
1	1	1	7	ABC

Fig. 63

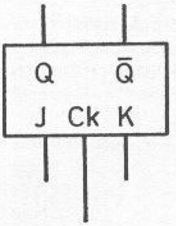
truth table. Each horizontal row corresponds to the state of the A, B and C outputs after an input pulse has been received, a logic 1 corresponding to a high point on the waveform, and a logic 0 to the low point on the waveform.

If A is given a decimal value of 1, B a value of 2 and C a value of 4, each of the horizontal lines corresponds to a decimal number.

In chapter 5, AND gates constructed from diodes and resistors were discussed. If a gate with three input diodes is connected to the outputs A, B and C, of the flip flops, the gate output can only be positive i.e. at a logic 1, when all three inputs are positive. The output of this gate is written as ABC in logic notation. The output waveform of such a gate is shown in Fig. 62, and it can be seen that it remains at a logic 1 for the time period between input pulses 7 and 8. Similarly another gate connected to A, $\bar{B}$ and $\bar{C}$ has an output function $A\bar{B}\bar{C}$ and corresponds to the time period between input pulses 1 and 2. Thus AND gates can be used to decode the binary counter outputs into decimal outputs. Fig. 63 shows the AND gate functions for the full count.

A counter and gate circuit of this type can therefore be used to produce separated signal pulses for each channel, by applying the received signal to the input of the counter, and taking the channel signals from the outputs of the AND gates. This type of decoder could easily be constructed from integrated circuits but is not very practical since it would occupy a printed circuit board area of approximately 4 square inches even if only four channels were required. The amount of current consummed would also be somewhat excessive. With discrete components a similar size problem would occur as well as the assembly costs being prohibitive from a manufacturing viewpoint.

There exists another type of flip flop called a JK. This has two inputs J and K which may be conditioned by continuous voltage levels, and a "clock" input to which the input pulse is applied. The final output after the clock pulse is determined by the combination of logic signals



J	K	Q_{n+1}
0	0	Qn
1	0	1
0	1	0
1	1	$\bar{Q}n$

Fig. 64

applied to the J and K inputs. Fig. 64 shows a block diagram of a JK flip flop, CK being the clock input. The truth table shows the conditions taken up by the outputs after the $(n+1)^{th}$ clock pulse has been received. Qn represents the state of the output prior to the $(n+1)^{th}$ clock pulse. It will be seen that with both the J and K inputs held at a logic 1 that the output takes up the complement of its state immediately prior to the $(n+1)^{th}$ clock pulse. With the J and K inputs both held at logic 0, the outputs remain unchanged.

The most useful aspect of the JK flip flop is that its output changes on the trailing edge of the clock pulse, but the J and K inputs can be changed after the leading edge has occured. This allows the flip flops to be connected as shown in Fig. 65. Consider all three flip flops to be reset to 0 state. The J input of the first flip flop will be at a logic 1 from $\bar{C}$, and the K input at logic 0. The other two flip flops have a logic 0 on the J input, and a logic 1 on K. Hence when the first clock pulse is applied, only the first flip flop changes state. If the reader were to follow through the sequence, it will be observed that only one flip flop

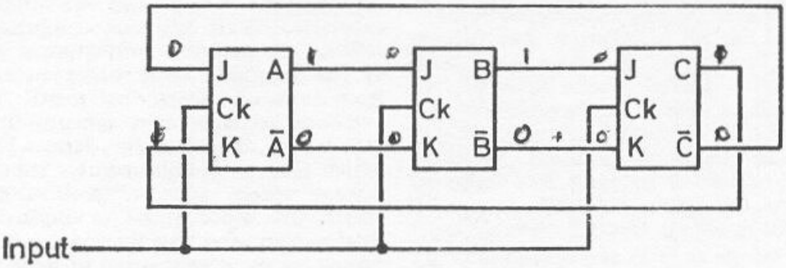


Fig. 65

changes state for each clock pulse. This has the advantage that no spurious spikes can occur from a decoded output, due to delays in flip flops changing state at different speeds.

Fig. 66 is a truth table for the twisted ring counter of Fig. 65. It will be seen that it is only necessary to use two inputs on each gate, since the gate functions shown only occur in one state of the counter. This form of decoder will provide decoding of five channels, the 000 state being used as a reset condition.

A	B	C	AND
0	0	0	$\bar{A}\bar{C}$
1	0	0	$A\bar{B}$
1	1	0	$B\bar{C}$
1	1	1	AC
0	1	1	$\bar{A}B$
0	0	1	$\bar{B}C$

Fig. 66

Both the binary counter and twisted ring are true counters in that they will continue counting for any incoming signals until reset. Thus a five channel twisted ring counter cannot be used with a six channel transmitter since channel six information would appear on the channel one output.

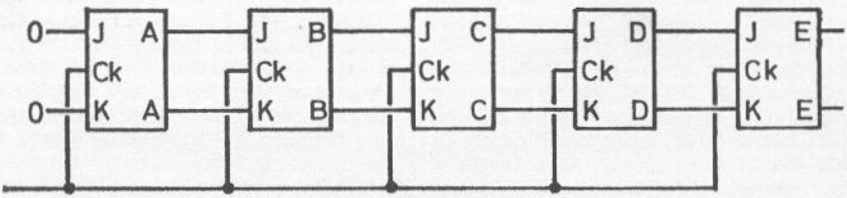
This disadvantage can be overcome by using a shift register as shown in Fig. 67.

If initially flip flop A is set to a logic 1 and the remaining flip flops to a logic 0, the truth table shown in Fig. 68 is produced. Any further input pulses applied will not produce erroneous outputs since only logic 0s will be propogated. This type of circuit has the disadvantage that a complete flip flop is required for each channel and a greater current consumption results compared with a counter and gate approach. Flip flop A can be used for a channel command by arranging for the synchronisation circuits to apply a logic 1 to the J input

A	B	C	D	E
1	0	0	0	0
0	1	0	0	0
0	0	1	0	0
0	0	0	1	0
0	0	0	0	1
0	0	0	0	0

Fig. 68

Fig. 67



and a logic 0 to K. The register then starts with all flip flops set to logic 0 and flip flop A becomes a logic 1 on the first clock pulse.

Synchronisation

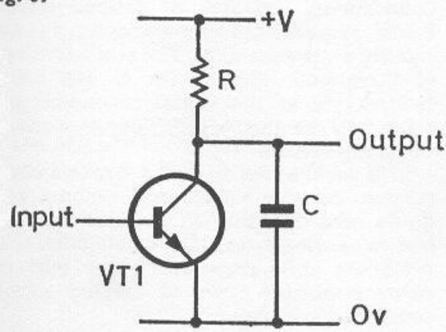
With the counter circuits described, it is necessary to ensure that each of the flip flops is in a reset state before the first pulse in the information is received. In chapter 6, when discussing encoding of the channel information, it was shown that provision for synchronisation was made by using a special long pause between pulses in the modulation signal (Fig. 32). Once a transmitter has been switched on, the first modulation pulse in the information frame is always preceded by this synchronisation pause. All that is required, is a circuit to detect the pause and apply a reset to the decoder counter whenever the pause is detected. This is readily performed by a capacitor charging up through a resistor, which is discharged by a transistor every time a modulation pulse occurs.

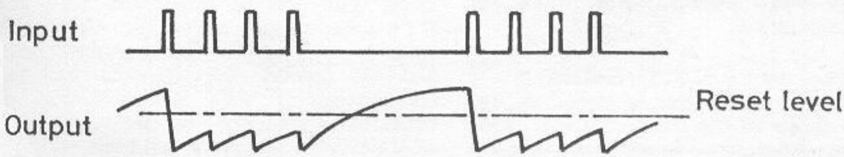
Fig. 69 shows an example of this type of circuit for supplying a positive reset, and is similar to the one used in the Microtrol system (Fig. 59). The incoming signal pulses cause VT1 to conduct with the result that C is rapidly discharged. When VT1 stops conducting, C charges towards +V with a time constant determined by the values of C and R. Before a reset can occur, the voltage on the capacitor must rise above a threshold determined by the characteristics of the decoder. For the circuit to operate corectly, it is a simple matter to choose the time constant so that the threshold is only exceeded during the synchronisation period. This is illustrated by the waveforms in Fig. 69. Most integrated circuit flip flops have provision for a Clear Input which overrides all the other input signals. Consequently there is no difficulty in causing a complete reset whatever the condition of the decoder flip flops.

Synchronisation of the shift register type of decoder operates slightly differently. The synchronisation pause is detected in the same manner as described above, but is not used to apply an overall reset. For the circuit of Fig. 67, if the inputs of flip flop A are held at logic zero, the first clock pulse sets A to logic 0, and this is propogated through until the fifth pulse is received. All stages are then reset and any further pulses maintain this reset condition. To allow the shift register to operate correctly as a decoder, a logic 1 must be applied to the J input of flip flop A until the first input clock pulse is received, and then removed for the remainder of the clock pulses. When a synchronisation pause is detected, the logic 1 must be re-established at the J input of flip flop A until the next clock pulse, this being the first pulse in the next information frame. Fig. 70 shows the input waveforms required to achieve this result.

Synchronisation of this form has the characteristic that if a decoder is switched on without a transmitter signal being received, it is possible for one decoder channel to remain on continuously. Re-

Fig. 69





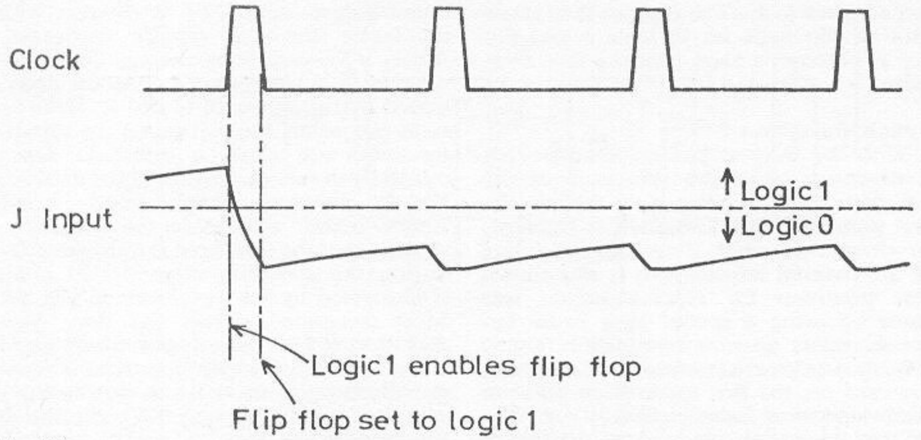


Fig. 70

setting of all stages in the counter type of decoder avoids this condition. This is not a particularly important point, since the type of servo amplifier employed has to be supplied with a pulsed input to produce motion. It has however been mentioned since it is one of the contributing factors affecting the degree of "kick" exhibited by servos when either the transmitter or receiver is switched on and off.

In Fig. 59, the synchronisation circuit for the Microtrol system comprises VT7 and associated components. During the synchronisation period, the collector of VT7 rises to a voltage higher than the input threshold of the first stage of the decoder shift register. On the leading edge of the next clock pulse, this logic 1 state enters the first shift register stage. C21 is rapidly discharged to 0v by VT7 conducting, but the collector voltage does not fall below the threshold level until after the propagation delay through the circuit. The consequence of this is that the first shift register stage becomes set only on the first clock pulse after every synchronisation period, as required for correct operation.

Transistor and S.C.S. Decoders

So far the discussion of decoders has been in terms of integrated circuits. This has been so described in order to avoid intermixing of circuit descriptions and logic functions. From a commercial viewpoint this may not necessarily represent the best approach for producing a decoder.

It is possible to design shift registers and counters using discrete components that will compete favourably in manufacturing cost with integrated circuits, with the added benefit of lower current consumption. Integrated circuits have the disadvantage that to make all the required connections, the area of printed circuit board required causes difficulty in designing a compact unit. The real advantage of integrated circuits lies in the lower failure rate at the initial manufacturing stage, and the greater reliability expectancy during operation.

It is uncommon to find a discrete component decoder employing counter and diode gate decoding in modern systems, but a large proportion of commercial units use shift registers. These may be either transistor types or employ silicon controlled switches (S.C.S.).

Fig. 71 shows a two stage shift register and synchronisation circuits, using transistors. The pair of transistors VT4 and VT5 form a latching circuit. One stable state is when both transistors conduct, and the second state is with both not conducting. If VT4 is conducting, its collector current becomes the base current of VT5. VT5 therefore conducts and part

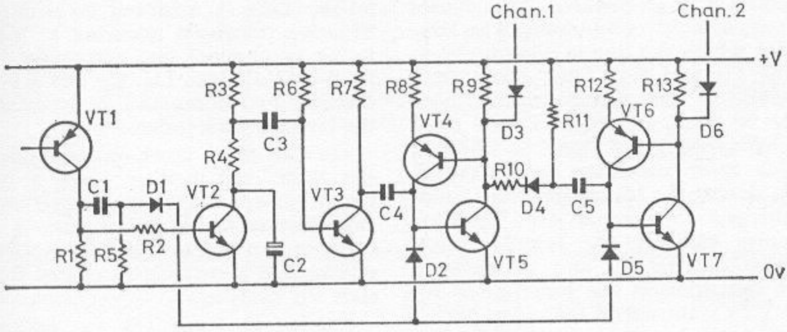
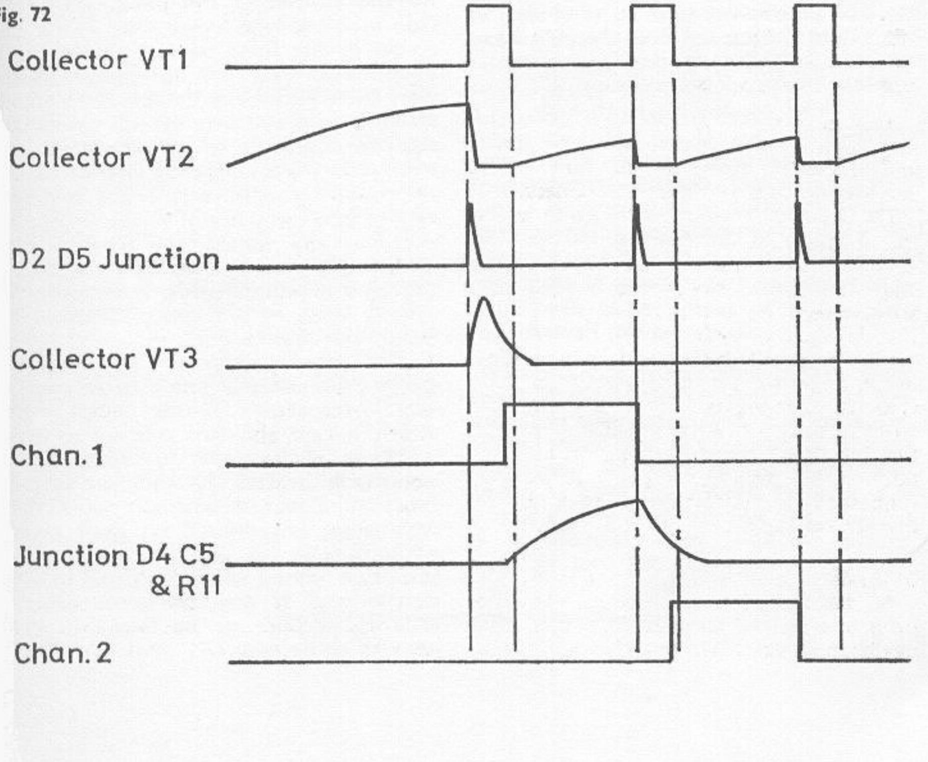


Fig. 71

of its collector current flows through the base of VT4 so maintaining this transistor in a state of conduction. If a negative pulse is applied to the base of VT5, it stops conducting and no base current can then flow through VT4 resulting in this also stopping conduction. A positive pulse on the base of VT5 has the opposite effect of causing both transistors to enter the conducting state. VT6 and VT7 form a similar latch for the second stage.

For a decoder application, a positive output is required on the channel lines and this corresponds to the appropriate stage being in the non-conducting state. Fig. 72 shows the waveforms through the circuit. A positive clock pulse at the collector of VT1, causes the common line

Fig. 72



to D2 and D5 to go positive. This causes both stages to start conducting. The time period for which the line is high, is determined by C1 and R5. Thus every clock pulse entering the system causes both stages to be reset. In order to set a stage a negative input pulse must be applied, after this reset condition has occured. Initially, during the synchronisation pause C2 is charged up towards +V. The first clock pulse entering the system causes VT2 to discharge C2 and produces a negative transition at the junction of R3 and R4. VT3 is turned off for a time period determined by the time constant C3 R6, chosen to be greater than the time constant of the reset line. As VT3 starts conducting its collector goes negative and this transition is coupled to the base of VT5 by C4. VT5 stops conducting, turning off the stage, so causing the channel output line to go high.

The next clock pulse entering the system causes the first stage to turn on again since a positive pulse is applied via D2. However, no negative pulse is applied since C2 has had insufficient time to re-charge to any extent during the first channel duration, and consequently the negative transition that appears at the collector of VT2 is of too small an amplitude to stop VT3 conducting.

To turn the second stage off and generate the second channel command, a negative transition must be applied to the base of VT7 immediately following the reset applied via D5 by the second clock pulse. This is achieved by slowing the negative transition occuring at VT5 collector as channel one command finishes, by means of R10, D4, C5 and R11. Thus channel two command is established by the second clock pulse.

On the third clock pulse, both stages are reset, but neither have a negative transition applied to turn them on again. Any further clock pulses have a similar effect until a synchronisation pause occurs, which then allows C2 to charge fully and start the cycle again.

Diodes D3 and D6 are included to prevent negative transitions present in the servo amplifier when a channel is operating, from turning on a stage that should be in the non conducting state.

The S.C.S. decoder operates in a similar manner to the transistor type described. The S.C.S. is effectively equivalent to the pair of transistors in each channel stage (Fig. 73) and differs only in that three connections are used whereas the transistor pair requires a separate connection for the emitter of the p.n.p. transistor. One problem with employing S.C.S. decoders is that these devices are primarily intended for operation on voltages higher than the 4.8v usually employed in digital systems. Consequently careful circuit design is required to accommodate the wide variation in operating characteristics, and some manufacturers found it necessary to grade batches of silicon controlled switches and adjust the load resistor accordingly. In spite of this, the S.C.S. proved a popular decoder component for several years before being displaced by integrated circuits.

The price of complex TTL integrated circuits has steadily fallen in relation to other components during recent years, and it is now standard practice to use a multistage shift register in a single encapsulation to provide the complete decoder function. The introduction of CMOS technology has offered an even greater advantage in the reduced current consumption offered when compared to TTL circuits and is now the most common type of decoder to be found in commercial equipment.

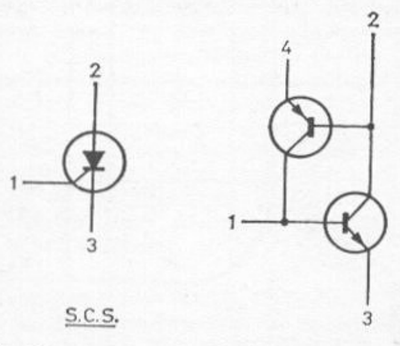
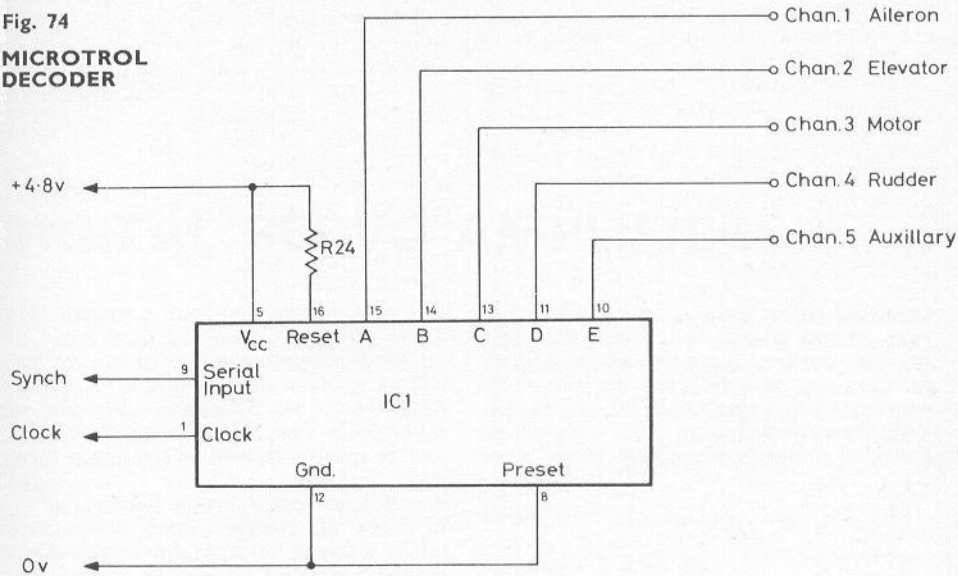


Fig. 73

Fig. 74

MICROTROL DECODER



Microtrol Decoder (Fig. 74)

The Microtrol Decoder comprises a five stage shift register integrated circuit, the SN7496. This is a multipurpose shift register and is used in 'serial in—parallel out' mode. Preset inputs to the SN7496 are not used in this application and these are tied to logic 1 and logic 0 respectively. The logical operation of the decoder is in accordance with the Truth Table of Fig. 68 but the actual circuit operation differs from Fig. 70 in that the outputs change state following the leading edge of the clock pulse. The only requirement that has to be satisfied in order to achieve correct operation is that the serial input to the SN7496 must be high at the leading edge of the first clock pulse following a synchronisation period, and must be below the input threshold for all other clock pulses.

The output stages of the SN7496 are suitable for driving the majority of servo amplifier circuits that require a positive output pulse and it is for this reason that a low power integrated circuit has not been used. With such a circuit the reduced output available would require an extra buffer stage for use with some servos. The extra capability of the integrated circuit used in the Microtrol system is achieved at the expense of slightly increased current consumption, but the magnitude of this is small compared with the overall consumption of a complete four channel system.

The decoder is mounted on the same circuit board as the receiver and the low profile of the integrated circuit allows ample room for a frequency change switch and fly leads to the servo connectors, whilst the overall receiver size is comparable with all but the smallest of commercial units without making assembly too difficult.

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CHAPTER 9

DIGITAL SERVO AMPLIFIERS

IN chapter 4 the basic components of a closed loop positional feedback servo were illustrated in block diagram form, by Fig. 17. The pulse-width tracking amplifier principle originated in the U.S.A., and whilst development has resulted in detail changes the overall system is unaltered.

The demand signal takes the form of the decoded channel pulse, the width of which defines the required position of the servo output arm. This pulse is repeated at the frame rate of the transmitter. Fig. 75 shows a block diagram of the servo amplifier system. The feedback generator takes the form of a pulse generator which is triggered by the leading edge of the incoming signal. The width of the pulse at the output of the generator is related to the position of the wiper of the feedback potentiometer. The output pulse therefore defines the present position of the servo output arm. The pulse is arranged to be of opposite polarity to the incoming signal so that a comparison can be performed at the summing junction. Fig. 76 illustrates the waveforms at this point in the system and shows how either a positive or negative error pulse is produced. This error signal defines the direction in which

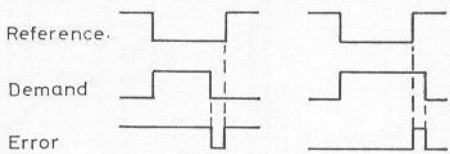
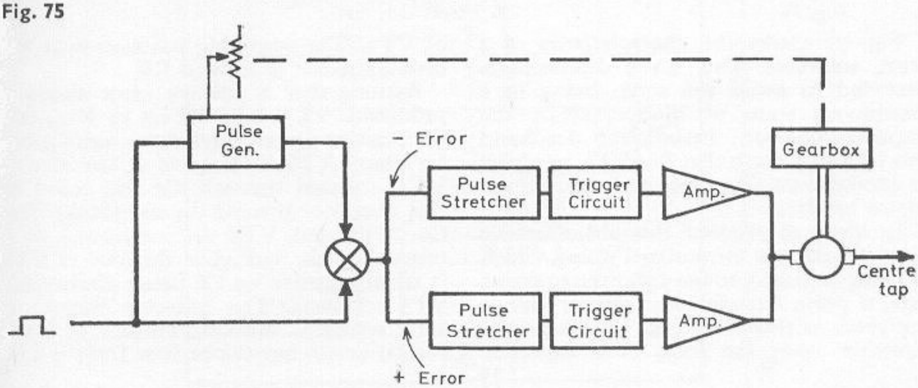


Fig. 76

the servo has to travel in order to take up the new required position.

If the channel signal varies between 1-2mS for the full range of servo travel and the servo is required to position within one percent, an error signal pulse of 10

Fig. 75



microseconds must supply sufficient input to the amplifier circuits to cause the motor to rotate. In order to overcome friction in the motor and gear train under conditions of small error signals, the drive signal to the motor takes the form of a pulse of full voltage amplitude. The mark-space ratio, or duty cycle, is increased from an initial value of around 50% until at approximately 10% error signal, a continuous voltage is applied to the motor. An advantage of this approach over the analogue method of applying a continuous voltage level, the amplitude of which is dependent upon the error signal, is that the power dissipation in the output transistors is minimised. With a pulsed drive, the transistors are either fully conducting, or completely turned off, so avoiding the intermediate conditions where power is dissipated.

which is discharged, and the time constants are arranged so that with errors in excess of 10% it is fully discharged. For smaller errors, the capacitor is only partially discharged. The trigger circuit operates for the time period that the capacitor is discharged below a threshold level, and so applies a voltage input to the drive amplifier for a time period dependent upon the discharge level of the capacitor.

Typical Circuit

Fig. 78 illustrates a high resolution amplifier following this approach. VT1 and VT2 and associated components form the pulse generator. RV1 is the feedback potentiometer, the wiper of which is mechanically linked to the output arm. The incoming channel pulse triggers the pulse generator via C2 and D1, and a negative output appears at the collector

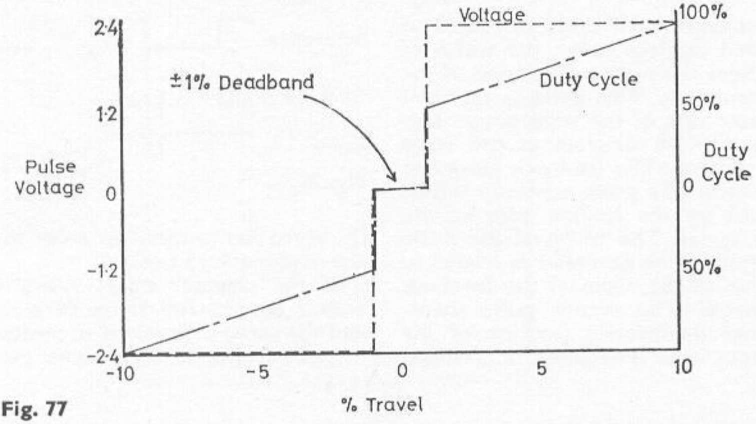


Fig. 77

Fig. 77 shows the characteristics of a servo amplifier. The ±1% deadband is provided to avoid the servo being in a continuous state of motion about the required position. Insufficient deadband can lead to a servo that "hums", resulting in excessive current drain and wear on the motor brushes.

In order to produce this characteristic from the narrow error signal pulse, which is being repeated at the transmitter frame rate, a pulse stretcher and trigger circuit are used, as shown in Fig. 75. The pulse stretcher takes the form of a capacitor

of VT2. The summing junction is at the common point of R7 and R8.

Assume that a positive error signal is produced. VT3 is biased off by R10, and the positive signal coupled through C3 has no effect. A positive signal at the base of VT4, coupled through C4 will result in this transistor turning on and discharging C6. VT6 and VT8 are connected as a trigger circuit, and when the base of VT6 is taken negative by C6 being discharged, VT8 conducts. The collector current of VT8 serves as the base current of VT9, and so causes current to flow from +4.8v

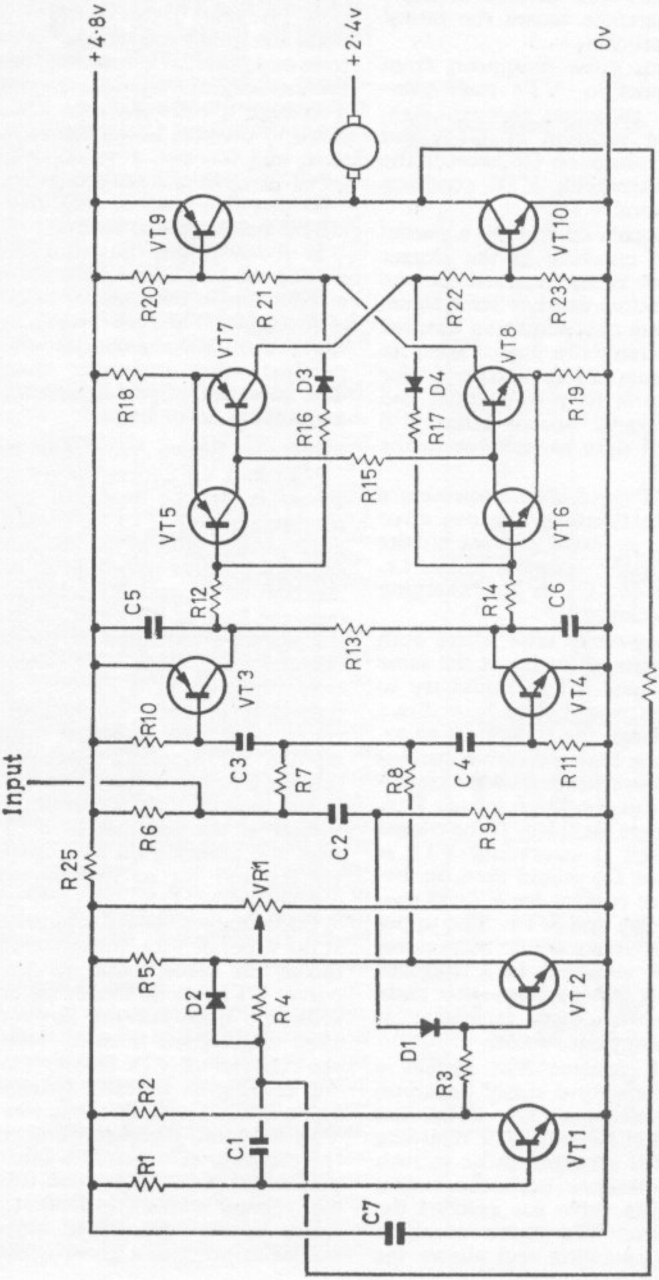

+4·8v
+2·4v
0v
Input
VT1
VT2
VT3
VT4
VT5
VT6
VT7
VT8
VT9
VT10
R1
R2
R3
R4
R5
R6
R7
R8
R9
R10
R11
R12
R13
R14
R15
R16
R17
R18
R19
R20
R 21
R22
R 23
R24
R 25
VR1
C1
C2
C3
C4
C5
C6
C7
D1
D2
D3
D4

Fig. 78

through VT9 and the motor, to 2.4v. This current therefore causes the motor to commence rotation.

When the error pulse disappears from the summing junction, VT4 stops conducting and C6 recharges through R13, R16, the emitter junction of VT5, and R18. When the voltage on C6 exceeds the trigger circuit threshold, VT6 conducts turning off VT8 and VT9.

Small error signals cause only a partial discharge of C6 resulting in the trigger circuit threshold being reached in the recharge cycle, before another error signal is produced by the next incoming channel pulse. This generates the pulsed drive to the motor for errors of less than 10%. For larger errors, C6 is fully discharged and next incoming signal occurs before C6 has had sufficient time to recharge to the trigger threshold.

VT3, VT5, VT7 and VT10 represent a similar circuit for handling negative error signals resulting in drive current to the motor flowing in the opposite sense, i.e. from +2.4v to 0v. C5 is the charging capacitor in this case.

If a condition were to arise where both trigger circuits were operative at the same time, both VT9 and VT10 would try to conduct. Current would then flow direct from +4.8v through the transistors to 0v. This would cause both excessive current drain, and rapid overheating of the transistors. To avoid this condition arising R16, D3 and R17, D4 are included. If the trigger circuit VT6, VT8 is operating, VT8 is conducting. Since D3 would then be forward biased, base current for VT5 is produced via R16, D3 and VT8. The upper circuit is thus maintained inoperative whilst the lower circuit is in a triggered state. R17 and D4 perform the same function when the servo amplifier is operating in the opposite sense.

The feedback resistor R24 applies a proportion of the output signal back into the reference pulse generator. This has the effect of either advancing or retarding the timing, and is arranged to be in such a sense as to produce no error output slightly before the servo has reached the required position. The drive amplifier therefore stops operating and allows the servo motor to coast to the required position. The value of this resistor is dependent upon the motor inertia, and the servo gear train and provides a means of compensating for mechanical variations between servos. Too high a resistor value will allow the servo to oscillate about the required position, and too low a value will make the servo sluggish and the resolution poor.

Whilst this form of amplifier offers excellent performance, with the miniaturisation of servos and the requirement of enclosing the amplifier within the servo case, a reduction in the number of components is desirable. This can be achieved at the expense of making the circuit more dependent upon transistor characteristics. The Microtrol circuit is typical of such an amplifier.

Microtrol Servo Amplifier. Fig. 79

VT1 and VT2 form the reference pulse generator. In the quiescent state VT1 is conducting and VT2 is turned off. The pulse leading edge of the incoming channel causes a positive transition to occur at the junction of C3 and R12. D2 couples this onto the base of VT2 and serves to isolate the negative transition at the end of the channel pulse, from VT2 base. The positive transition at VT2 base causes this transition to start conducting with the result that its collector voltage falls rapidly. C2 therefore discharges via D1 causing a negative transition to be applied to the base of VT1 so turning it off. The voltage at the collector of VT1 therefore rises and allows VT2 base current to flow via R3 and R4 so maintaining VT2 in conduction.

C2 recharges via R5 towards the voltage at the wiper of VR1, the feedback potentiometer. At some point on the recharge curve, VT1 can no longer be held off and sufficient base current flows via R1 to allow conduction to start. The voltage at the collector of VT1 falls turning off VT2 and causing its collector voltage to return to a positive level. The collector voltage of VT2 therefore represents a negative pulse, the duration of which is a function of the pot. wiper position. R7 and R8 determine the voltage across the feedback pot. and hence specify the wiper movement required to produce a given voltage change.

Since this voltage in turn determines the reference pulse width, R7 and R8 determine the range of mechanical movement for a given range of input signal.

R11 serves as a current path to capacitors C3, C4 and C5, and also represents the terminating resistor for the diode gates of the decoder. The junction of C4, C5, R10 and R13 is the summing junction, at which point the widths of the incoming pulse and reference pulse are compared. The error pulse coupled via C4 or C5 causes either VT3 or VT4 to conduct, dependent upon the sense of the error. In quiescent state, C6 is charged to +2.4v via R16 and R19 and neither VT5 or VT6 conducts.

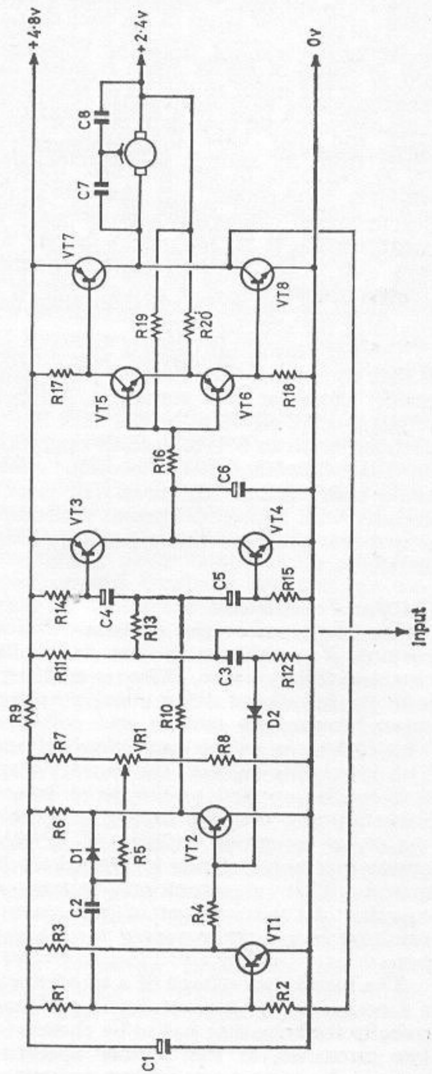
If VT3 is turned on by the error signal, C6 is rapidly charged to its collector voltage (approximately +4.5v) so allowing VT5 to conduct. In this condition VT7 conducts causing current to flow from +4.8v through the motor to +2.4v. VT5 base current flowing from +2.4v via R20, through VT5 and R16, causes C6 to discharge back towards +2.4v eventually turning off VT5 and VT7.

For a positive error signal, C6 would be discharged below +2.4v causing VT6 and VT8 to conduct until it recharged to +2.4v. It should be noted that only half of the amplifier can be on at any time since the pulse stretching capacitor C6 is common to both. This eliminates the cross clamping components needed in the previous circuit. The omission of the trigger circuits requires high gain transistors to be used for VT5 and VT6 otherwise insufficient base current will be supplied to VT7 and VT8. VT5 and VT6 must also be of near identical characteristics otherwise the servo will respond differently in each direction of travel.

Feedback to the reference pulse generator is provided via R2 and has to be applied at a different point to that in Fig. 78. This is because the sense of the output signal is inverted due to the omission of one transistor stage in the simplified amplifier. C1 and R9 provide decoupling of the reference pulse generator to avoid spurious triggering due to noise on the supply lines casued by the servo motor. C7 and C8 are included for radio frequency inter-

Fig. 79

MICROTROL SERVO AMPLIFIER



ference suppression purposes.

Input Buffers

Both the circuits illustrated require the decoder output to be able to provide sufficient current to both trigger the pulse generator, and operate the summing junction. Whilst this is the case with the integrated circuits of the Microtrol decoder, some other types of decoder cannot provide the necessary drive. To overcome this problem, a buffer stage is often included at the input to the servo amplifier. This takes the form of a single transistor stage and may have the output taken either from its emitter of collector, dependent upon whether it is required to invert the channel signal or not. Some S.C.S. types of decoder have a negative output and the invertion is used to produce a positive input to the servo. The remainder of the servo amplifier then becomes of the standard type.

Another reason for using a buffer stage is that without it, the reference pulse generator signal at the summing junction causes a small signal to be fed back to the decoder. With an S.C.S. decoder operating on low currents, this transition could cause malfunction. An integrated circuit decoder with its higher current consumption is less prone to this type of external interference.

Output Transistors

The choice of output transistor is important if maximum power is to be obtained from a servo. When a transistor is in its fully conducting state, a voltage exists between its emitter and collector. This is known as the saturation voltage. The lower this voltage, the more voltage that can be applied to the servo motor. Remembering that the motor runs from +2.4v, a saturation voltage of 0.7volts means that only 1.7volts is applied to the motor. With a saturation voltage of 0.3volts, 2.1volts is across the motor, resulting in greatly increased torque output.

The saturation voltage of a transistor is a function of the current flowing. Consequently the transistor has to be chosen for low saturation at the normal operating current. In general, germanium transistors have a lower saturation voltage than silicon, and for this reason many early amplifier designs used germanium types in the output stages. Improved manufacturing techniques have produced silicon transistors with reduced saturation voltage and these are now almost universally used, their plastic encapsulation offering the advantage of reduced physical size.

The power rating of the transistor is important since it must be able to cope with a stalled motor. With a 5ohm motor in a stalled condition, operating on 2.4volts, and a transistor with a saturation voltage of 0.7v, the current flowing is 340mA. The transistor is required to dissipate $0.7 \times 340 = 238$milliwatts. For 0.3v saturation voltage, the current is 420mA giving a power dissipation of 126milliwatts. A 2.5ohm motor will double the currents and hence the power dissipated. The majority of suitable silicon planar transistors will dissipate a maximum of 400milliwatts. Hence for the low resistance motor it is imperative that a low saturation voltage transistor is used, and for the 5ohm motor it is desirable. The saturation voltage must apply at the motor stall current for these purposes.

In operation, a pulse of current approaching the value of the stall current occurs when the motor is in a stationary state. Once in motion, the current is greatly reduced, and provided the switching is clean, there should be no tendency for output transistors to run other than slightly warm.

Voltage and Temperature Stability

The positional accuracy of a servo is dependent upon the accuracy with which the potentiometer and reference generator circuit convert the output arm position into a pulse. It can be shown mathematically that provided the transistors in this circuit saturate to a very low voltage, the circuit becomes virtually independent of supply voltage variations. In a similar manner to the transmitter encoder, refer to chapter 6, the output pulse width is dependent upon the ratio of the potentiometer resistance "above" the wiper, to the resistance "below".

A higher saturation voltage of the transistors will result in the reference

generator pulse output for a fixed wiper position, exhibiting a tendency to vary with supply voltage. With the circuits of Fig. 78 and 79, the variation between a fully charged 4.8v nickel cadmium cell, and one discharged to its minimum useful level, results in a servo positional drift of about 6%. This can be eliminated by using a zener diode to voltage stabilise the reference generator, but at the expense of increased current consumption.

In normal usage, both the transmitter and receiver system batteries are charged together, and during operation their voltages both fall. Hence both the transmitter encoder pulse and the reference generator pulse, drift in the same sense. Thus the servo drift occuring in practice can be expected to be far less than 6%. Also since the voltage of freshly charged batteries falls rapidly to a lower level and then tends to stabilise, the majority of drift occurs during the first few minutes of operation. Hence if a practice is made of switching on the transmitter and receiving system for a few minutes after recharging, the resultant drift during operation will be almost negligible.

Another factor that can affect accuracy, is temperature stability of the timing components. A disc ceramic capacitor used as C2 in Fig. 79 would result in considerable servo drift over a normal ambient temperature range. It is therefore usual to employ plastic film capacitors at this critical point, as these have a better temperature coefficient. The associate timing resistor R1 does not need to be of a special type, since the temperature coefficient of resistors is several orders better than that of most types of capacitor.

Integrated Circuits

The use of standard integrated circuits in servo amplifiers is somewhat restricted, since the nature of the circuit is considerably different to that normally encountered in most branches of digital electronics. As a consequence an attempt to use integrated circuits often results in a circuit which is more complex than it discrete component equivalent.

The first exception to this is the use of a resistor-transistor logic element, in an encapsulation similar to a medium sized transistor, for the transistor and associated resistors in the reference pulse generator. The disadvantage of this is that the application is using the integrated circuit in a manner for which it was not originally intended. As a result the characteristics that are important in this application are not those which are tightly controlled during manufacture. Consequently a wide variation in performance can result from nominally identical circuits. Also since the timing resistor is incorporated in the integrated circuit and is of comparatively low value, the associated capacitor becomes of large capacity. To maintain small physical size a tantalum type has to be employed and the temperature stability of these is only marginally sufficient for the accuracy required. It is probably fair comment to say that the use of this type of servo amplifier achieves miniaturisation at a sacrifice of performance. However, for the majority of uses of proportional radio control, the performance is more than adequate.

The majority of manufacturers now employ custom designed integrated circuits. These comprise a reference generator, summing junction, pulse stretcher, error amplifiers and at least part of the output stage, all in a single encapsulation. The timing components, pulse stretcher capacitor and feedback resistors are external to the encapsulation, so allowing the servo manufacturer control over these critical factors.

By incorporating additional transistors in the output stage and eliminating the battery centre tap, the resultant bridge output circuit has a lower power dissipation in the transistors so allowing inclusion of part of the circuit in the integrated module. Manufacturing techniques preclude the inclusion of both PNP and NPN transistors in the same integrated circuit and so it is usual to require two PNP transistors external to the encapsulation.

The following table lists the basic characteristics of servo amplifiers commonly used by UK manufacturers.

Type Number	ZN403E	SRC419	NE543	NE544	SN28604
Manufacturer	Ferranti	Skyleader/ Ferranti	Signetics	Signetics	Texas
Input Pulse	Positive	419P Positive	Positive	Positive	Positive
3-4 Wire	4	3	3	3	3
Pot. Resistance (Ohms) ...	1·5K	1·5K	5K	5K	5K
Motor Impedance (Ohms) ...	3-5	11	11	11	11
Supply Voltage (Volts) ...	3·5-6·5	3·5-6·5	4-6	3·5-6	3-6
Package	14 pin DIL	14 pin DIL	10 pin round	14 pin DIL	12 pin DIL
Output Circuit ...	2 external transistors Centre tapped supply	Bridge circuit requiring 2 external transistors	Bridge circuit using NPN transistors all in chip.	Bridge circuit using NPN transistors all in chip.	Bridge circuit using NPN transistors all in chip.

Note:- This table represents types generally available in UK at the time of publication, but is not intended to be an exhaustive listing.

Microtrol I.C. Servo Amplifier (Fig. 79a)

The Microtrol integrated circuit servo amplifier uses the SRC 419 circuit module together with two PNP transistors and is compatible with most modern receivers giving a positive pulse decoder output.

C2 provides a d.c. block on the input signal and current is limited by R1. The reference generator pulse width is determined by R2 and C5 and is modified by the voltage on the potentiometer wiper RV1. R4 causes the wiper of RV1 to be positioned near the centre of the track. C6 is decoupling to prevent strong transmitter signals from being picked up by the timing circuits. The pulse stretching function is controlled by R3 and C4 and the servo deadband is set by C3. VT1 and VT2 are PNP transistors complementing two NPN types included in the encapsulation, each complementary pair driving the motor in one direction only. R5, R6, R7 provide feedback controlling the dynamic damping of the servo. Adjustment of R6 and R7 allows selection of the desired amount of overshoot for the servo mechanism. C1 provides supply decoupling whilst C7 and C8 are suppression capacitors for the servo motor.

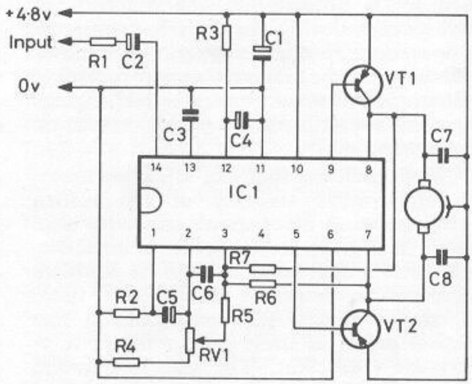


Fig. 79a

MICROTROL I.C. SERVO AMPLIFIER

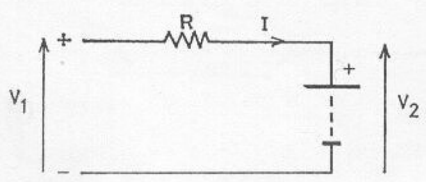
CHAPTER 10

CHARGERS

THE majority of commercial equipment with more than two servos employs rechargeable batteries for both the transmitter and receiver. The most common available in the U.K. have been those manufactured by DEAC, but recently other manufacturers' products have entered the market. These are all of the nickel cadmium type since with an appropriate type of physical construction, they are capable of delivering heavy currents for short periods of time. A word of warning at this point, DEAC manufacture two types of cell that are externally identical. The DK type are intended for supplying low currents for long periods whilst the DKZ type are capable of delivering heavy currents without suffering from a large internal voltage drop. Only the DKZ type are suitable for radio control applications.

For all these cells the recharging requirement is that they be recharged for a minimum of 14 hours at the 10 hour rate. Thus a 500mAHr cell should be recharged at 50mA for 14 hours. No damage will result if the charging current is applied for longer than this time. It is possible to recharge at a higher rate for a shorter period, but once the cell is fully charged, if the charging current is not reduced, irrepairable damage will result. Since it is generally not possible to know the extent of the charge remaining in a cell, high recharging currents should be avoided completely.

Fig. 80



The requirement of a simple charging circuit is illustrated in Fig. 80. The discharged cell has a voltage V_2 across its terminals and the charger supplies a voltage V_1 that is greater than V_2. Hence current flow occurs from V_1 to V_2 through the resistor R. The charging current is therefore

$$\frac{V_1 - V_2}{R} \text{ Amps.}$$

Since the value of V_2 increases slightly as charging progresses, the charging current will be reduced. Thus a charger circuit can be designed using the nominal cell voltage, resulting in a higher current for a discharged cell, and a lower one for a charged cell. There is therefore no danger of exceeding the charging restrictions on the cell.

Voltage Reduction

The low value d.c. voltage V_1 has to be derived from the a.c. mains which for the most of Europe is around 230v and in the U.S.A. is 115v. Some means of voltage reduction followed by rectification is therefore required. It is possible to use a capacitor for voltage dropping, but suitable types capable of handling the voltage are costly. A capacitor breakdown in this application is likely to result in an excessive charging current with disastrous results. A failure of a resistive mains dropper is not likely to be so dangerous

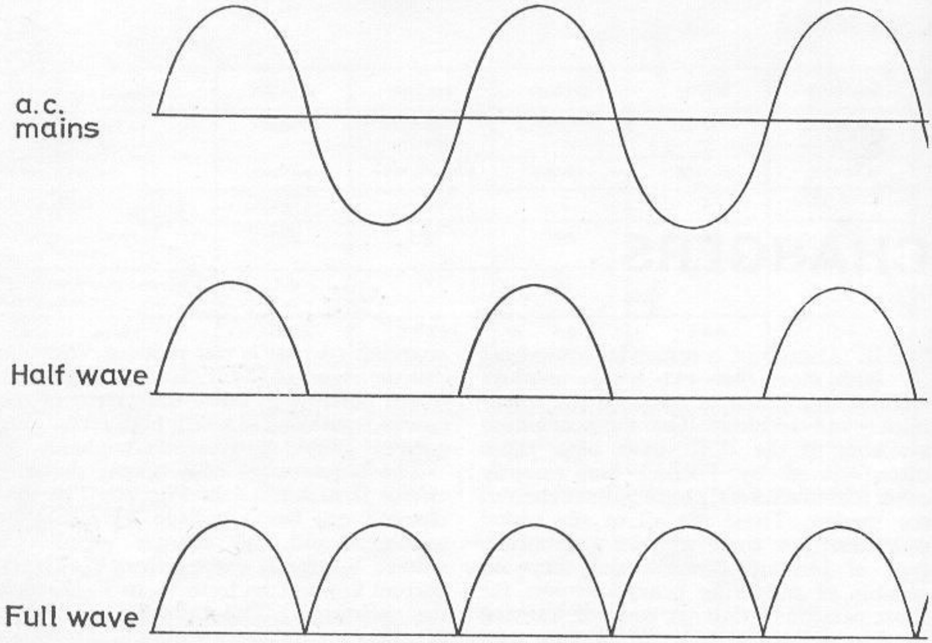


Fig. 81

since should the resistor breakdown, the charging current is likely to be reduced.

The disadvantage of a resistor in this application is that heat will be generated since the resistor will be consuming power. With the lower voltage of the U.S.A. mains supply, this heating is tolerable. However, the European mains voltage results in a fourfold increase in heat dissipation, so this form of charger is not very common.

By far the most satisfactory solution is to use a voltage stepdown transformer to produce a voltage in the same order as that of the cell to be charged. This has the added advantage that it provides a greater degree of isolation from the mains than does either of the other two methods.

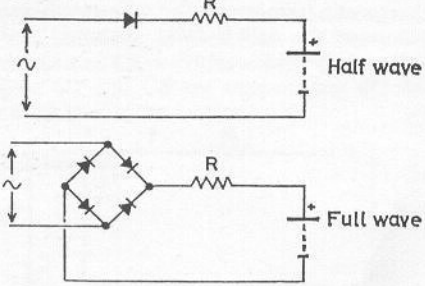
Rectification

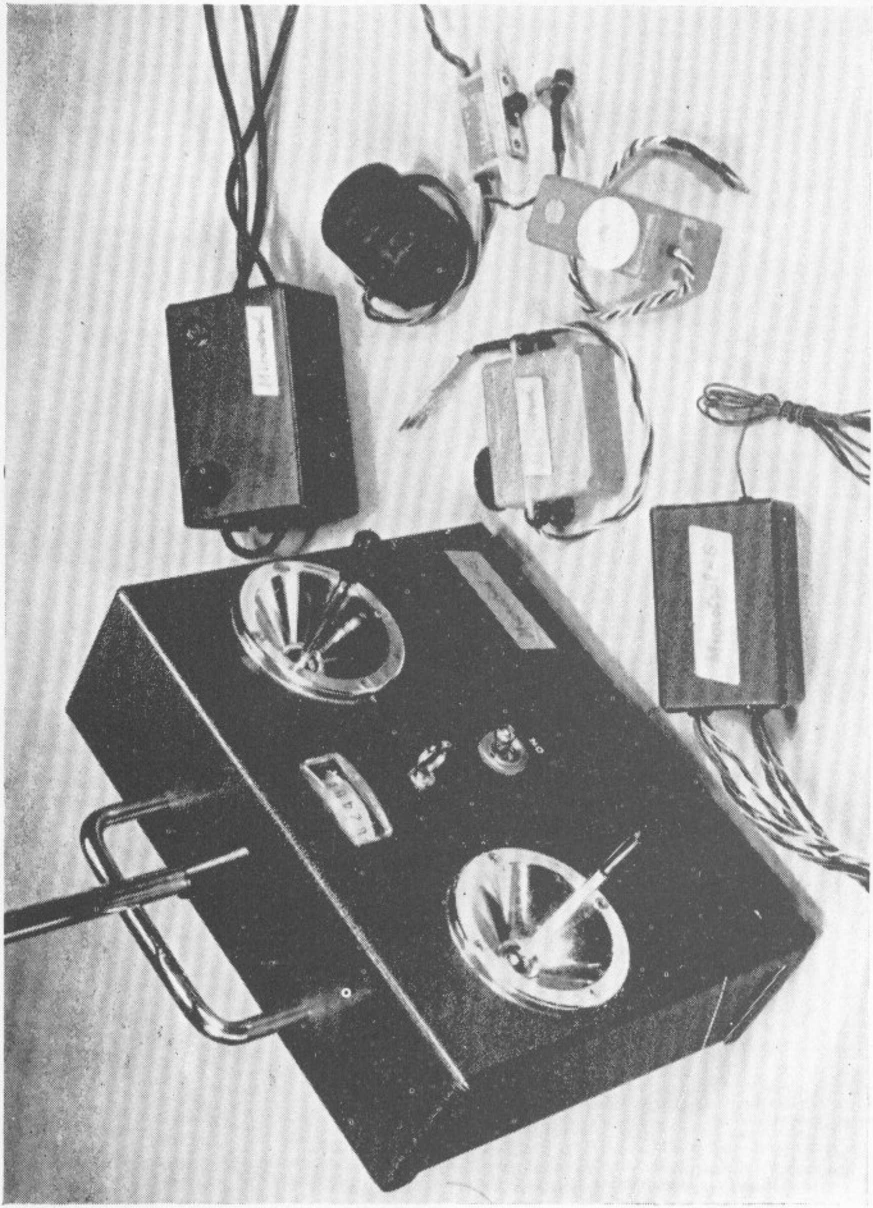
To produce a d.c. voltage from an a.c. supply, two forms of rectification may be employed; half wave and full wave. Fig. 81 shows the waveforms for both these methods. The effective d.c. voltage of the output waveforms from a rectifier, is a function of the average value of the waveform. Clearly the full wave rectifier will provide a higher d.c. voltage.

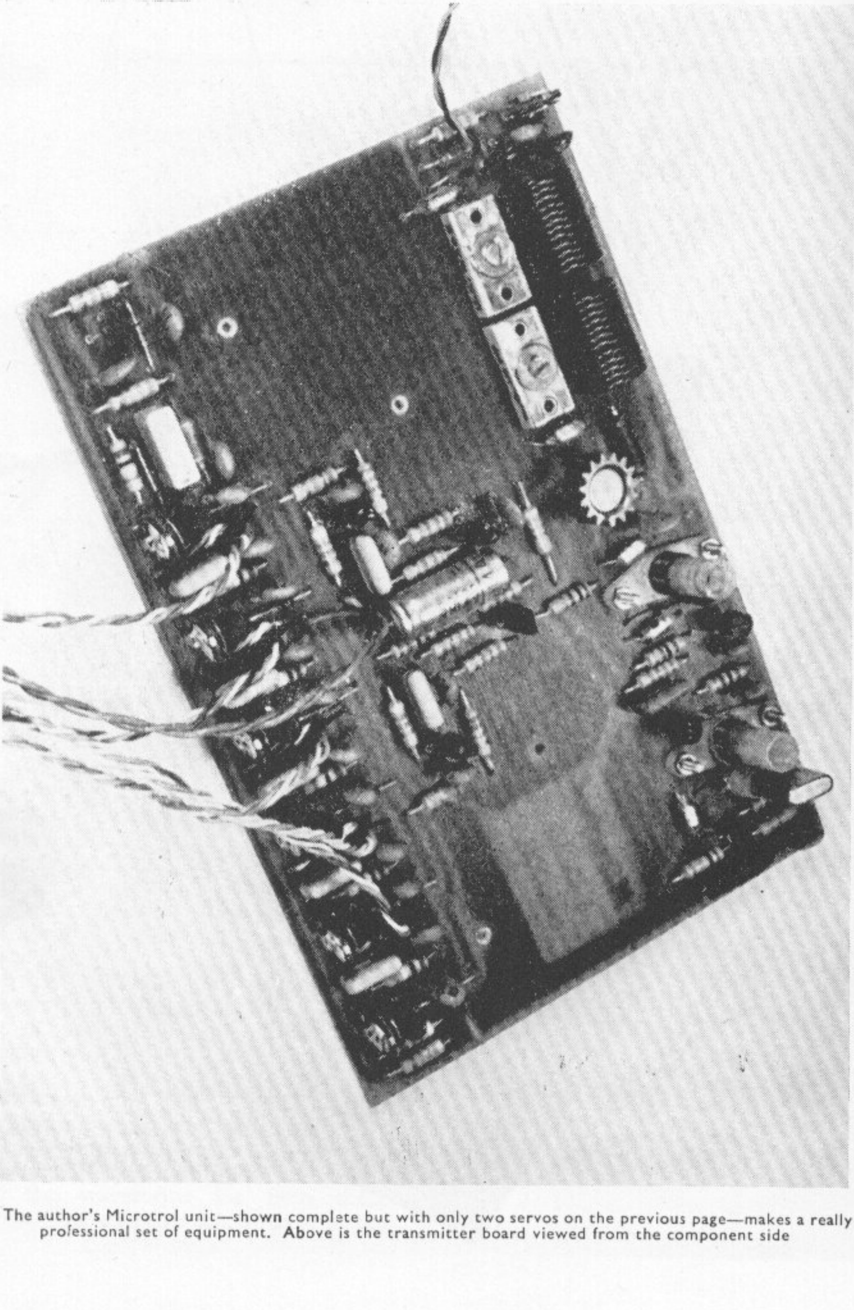
Either form of rectification is suitable for a charger, the only difference being that the transformer secondary will require a higher nominal output voltage if half wave rectification is used.

Fig. 82 shows a typical charger circuit for each type of rectifier. The value of

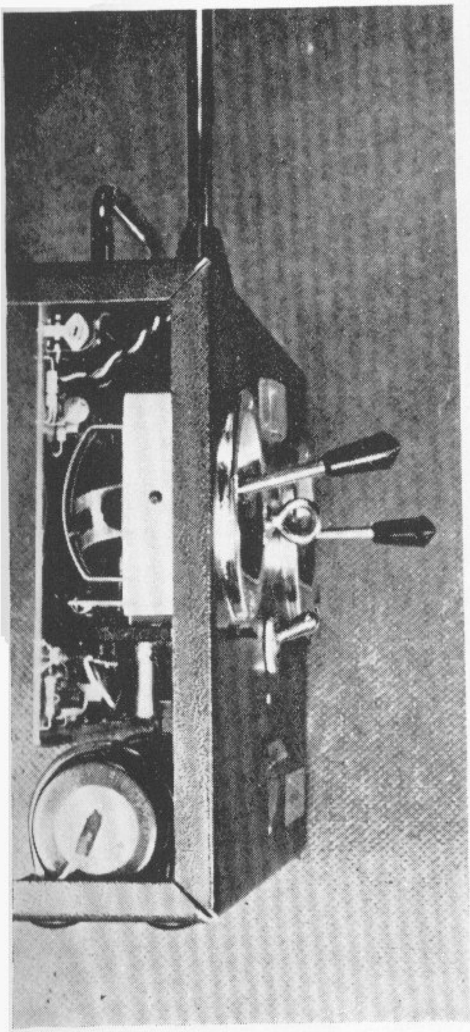
Fig. 82

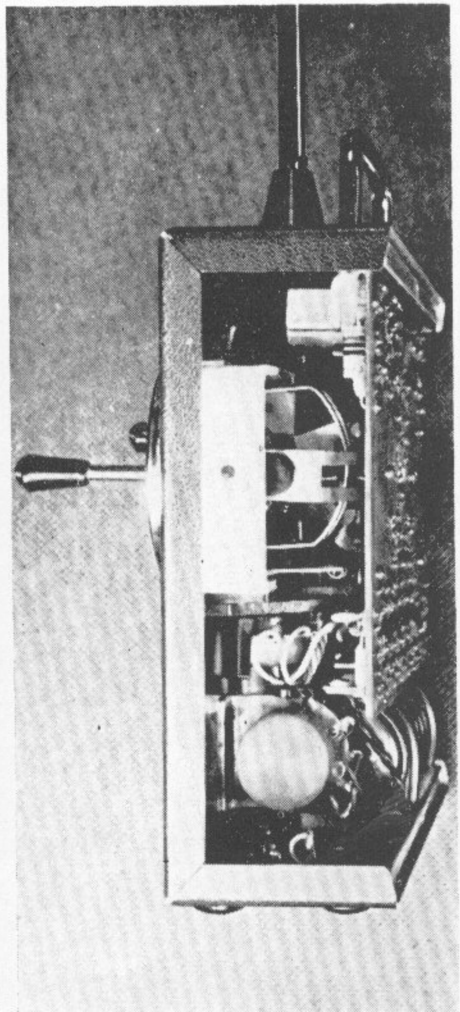




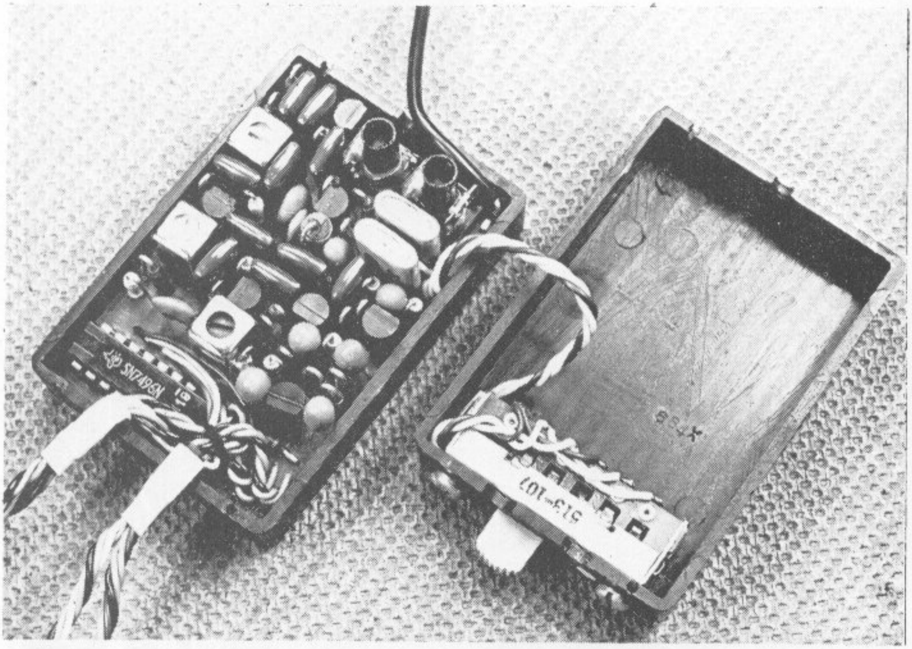


The author's Microtrol unit—shown complete but with only two servos on the previous page—makes a really professional set of equipment. Above is the transmitter board viewed from the component side

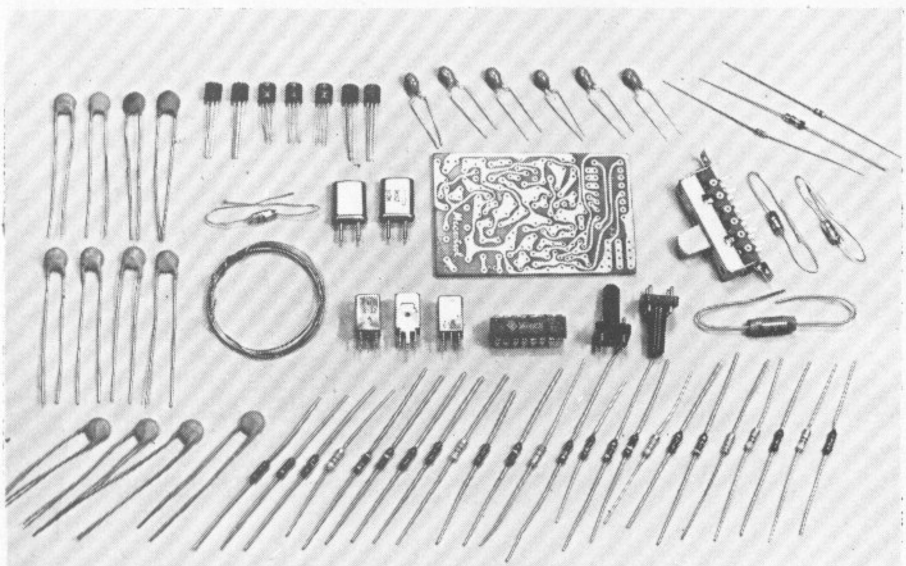


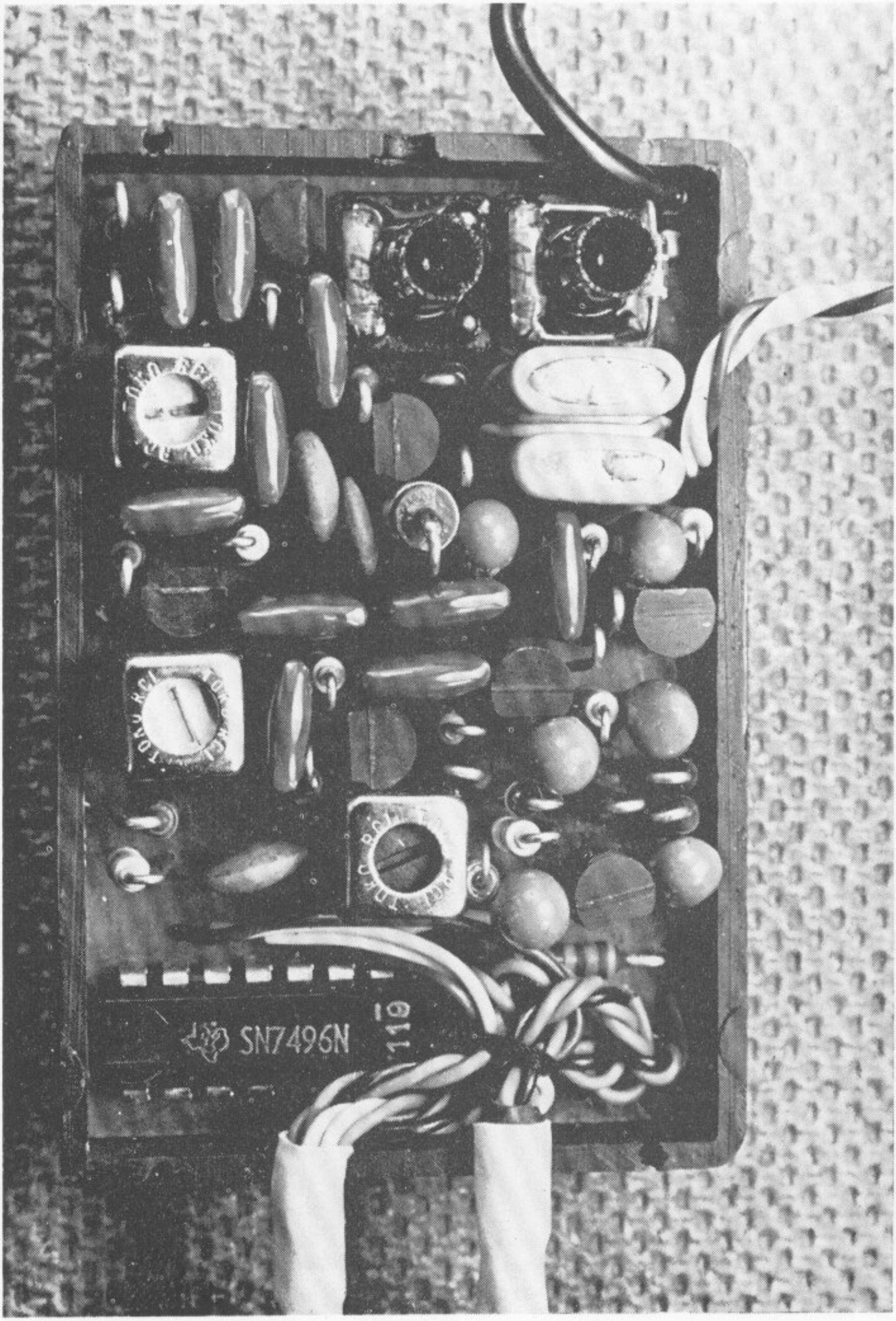


Two views of the transmitter from the side—note the compact arrangement and the position of the auxiliary control pot. on this prototype version



Finished receiver and decoder in case about actual size, components being shown below. Photo opposite shows enlargement of the complete receiver

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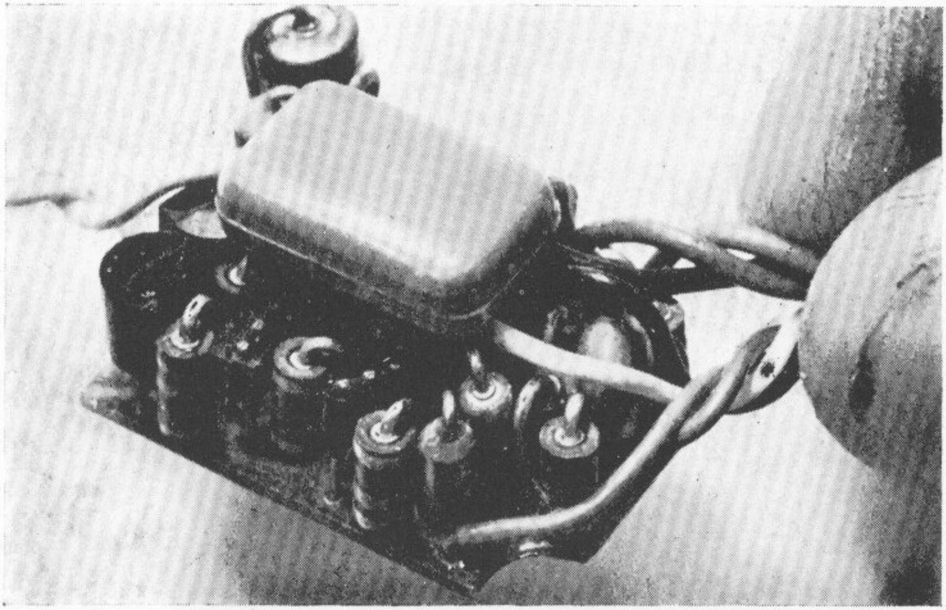
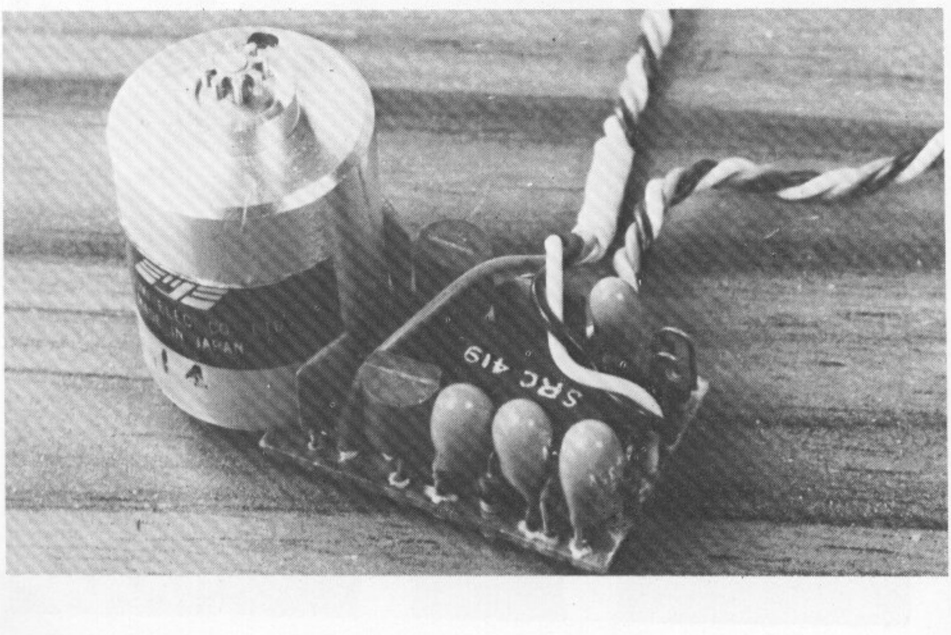
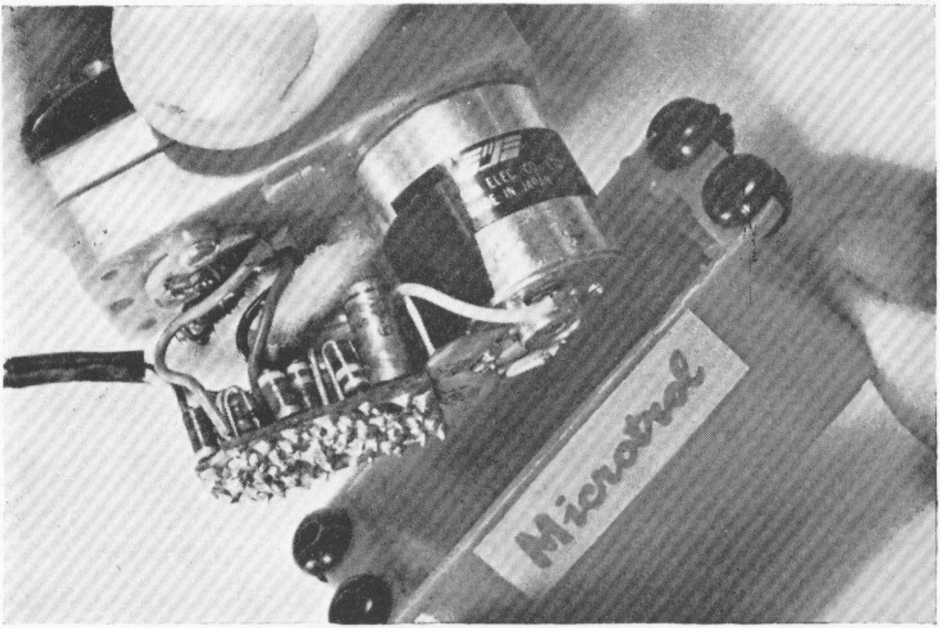
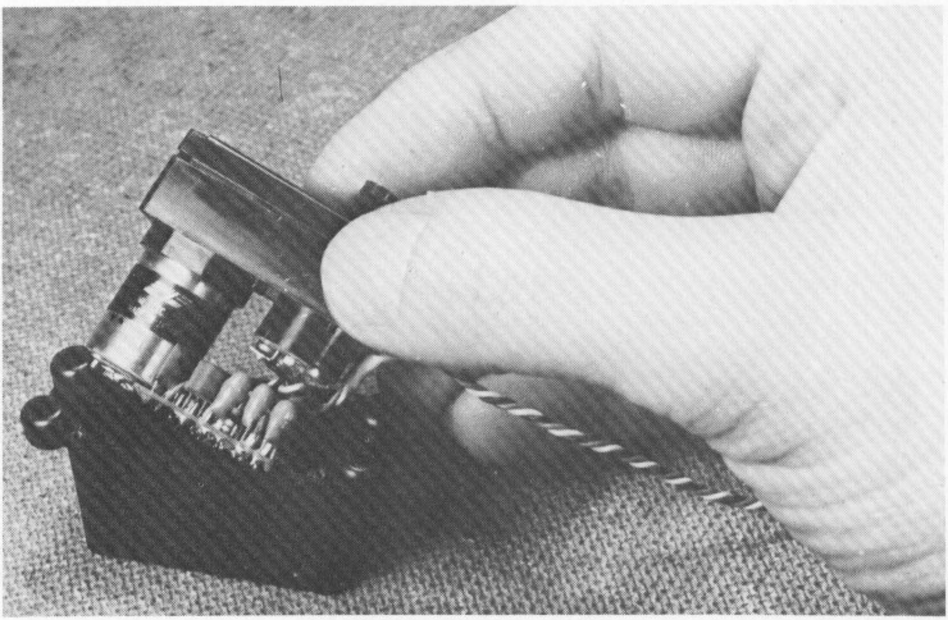


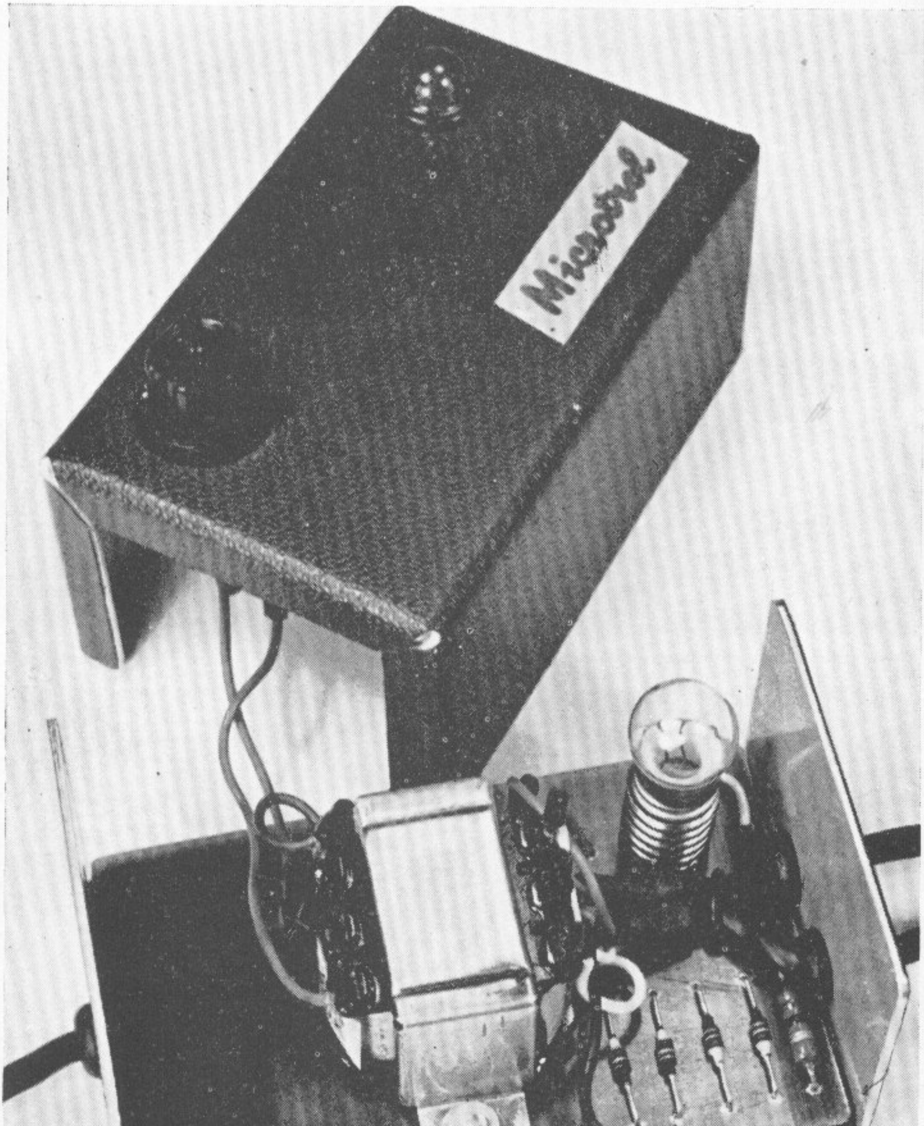
Top illustrations show the descrete amplifier after assembly and being fitted into an S.L.M. FB-3 mechanics unit. Most popular types of servo mechanics may be substituted.

Below: The i.c. amplifier complete with motor soldered in place, and being introduced into an SRC-1 mechanics unit.




Microtrol





The Microtrol charger unit. Neat construction is typical of the author's professional approach to his " home-built," justified by many hours trouble-free flying

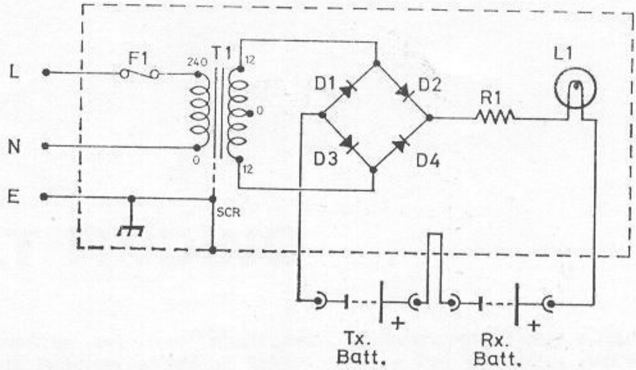


Fig. 83 MICROTROL CHARGER

R is chosen to provide the correct charging current and may be made adjustable if required. Should this be done, R should comprise a variable resistor in series with a fixed resistor, so that the maximum charging rate cannot be increased beyond the safe maximum for the cell.

Multiple Outputs

A charger can be designed with outputs for charging a range of different cell voltages. To achieve this, the transformer is chosen to provide enough voltage for the highest voltage cell, and separate resistor is provided for each of the outputs. In either of the circuits of Fig. 82, the extra resistors would be connected to the diode(s) at the same point as the existing one.

Alternatively if a charger is required to charge two batteries, e.g. the transmitter and receiver batteries, they can be connected in series and charged as a single unit. This approach is particularly common where a charger is included in the transmitter.

Battery Conditioners

It is an established fact that nickel cadmium cells exhibit a loss of capacity when repeatedly cycled through a partial discharge and recharge process. To obtain maximum performance from such a battery it is desirable to reach the fully discharged state before commencing charging. A battery conditioner achieves this by applying a load to a partially discharged battery, monitoring the battery voltage to detect when the fully discharged voltage is reached, and removing the load at this time. A normal charging cycle is then entered.

Repeated cycling in this sequence will restore a normal battery to its full capacity capability. The same effect can be produced manually by the use of a dummy load and monitoring the battery voltage with a test meter. Care is required to avoid discharging the battery below about 1.1 volts per cell otherwise damage may be caused.

The Microtrol Charger (Fig. 83)

This is a full wave charger for recharging the transmitter and receive batteries of the Microtrol system. The 9.6v and 4.8v batteries are connected in series by means of fly leads emerging from the p.c. board.

The charger incorporates a fuse on the mains input and a lamp bulb to indicate that charging is taking place. The lamp bulb is so chosen that should a cell be faulty resulting in excessive current being taken, the bulb acts as a fuse and blows. A series resistor, in conjunction with the bulb, limits the charging current to 50mA for the 500mAHr batteries.

CHAPTER 11

FAULT FINDING

FAULT finding on any electronic equipment requires specialist knowledge, and this includes radio control equipment. The reader who has studied all of the preceding chapters should not consider himself qualified even to take a screwdriver to a piece of commercial equipment. The best repair is that performed by the person who knows the equipment best, and this in every case is undoubtedly the manufacturer. We have often had opportunity to examine equipment returned to the manufacturers after having been investigated by "a friend in electronics", and can only say that we hope that we never have the misfortune to purchase some electronic equipment with which the "friend" has been concerned! No manufacturer is going to thank a customer who has attempted to "help" in finding the fault. The only point we would stress is that before returning the equipment, the customer is certain that the fault does not lie with the operator or with the model itself. Another book in this series, the "Radio Control Guide" has a chapter on fault finding from the viewpoint of a malfunction during operation, and it is not proposed to repeat the same information.

The constructor of kit equipment is best advised to follow the instructions to the letter, and if the equipment does not perform exactly how it should during the test procedure, then it should be returned to the manufacturer.

Test Equipment

There is a different requirement for the equipment required to check out an outfit, to that required to fault find. Many kitted systems require only a test meter to carry out a full alignment process. However if there is a fault, with the complexity of proportional equipment, an oscilloscope is invariably required. This allows waveforms through the system to be examined, and to the experienced operator, who understands the functioning of the digital outfit under test, it becomes a simple matter to follow the signal through up to the point where malfunctioning is present.

One little point that we learned very early on, is that any test equipment leads whether test meter, oscilloscope or digital timer, must have a resistor of at least 10K in each lead at the test clip end. If this is not done, the effect of a long length of wire attached to either a transmitter or receiver p.c. board is to upset the tuning, due to this lead acting as a dummy earth. Allied to this, any mains operated equipment should not have either test lead connected directly through to the earth pin on the mains plug. If this were the case, again the tuning would be disturbed since the mains earth would be attached to the p.c. board.

Investigating Faults

Having stressed the point of not fiddling with commercial equipment, there will still be a group of experimenters or home builders who would like some guidance on fault finding. We will assume that the equipment in question is known to be a working design as distinct from a new "home brew" still in its initial stages. This is necessary since it is beyond the scope of this book to cover design faults, only faults caused by some form of failure will be considered.

Solid faults are obviously easier to find than intermittent ones. The cause of a once only "glitch" will probably never be found, but one that occurs under fairly repeatable operating conditions can usually be found on the work bench.

The first consideration is to identify the nature of the fault and then to localise it. If a fault occurs on one servo, it can be either in the transmitter encoder, the decoder, or the servo. Where servos are interchangeable it is a simple matter to identify whether the servo is at fault. If the outcome of this points to the encoder or decoder it is worth examining the transmitter output on an oscilloscope before attempting to test within the system. An encoder fault will almost certainly be evident in the output waveform. If not, then the decoder should be investigated. It is at this point that a detailed knowledge of the system is required since it is necessary to examine the operation of the decoder in a logical manner. No general guide lines can be given other than to follow the signal through from the beginning.

A symptom common to all servos can be caused by a legion of faults. Always check the battery voltage under operating conditions first. Low voltage can cause strange symptoms and it is all too easy to spend time investigating one of these. Assuming the batteries are good, the next stage is to determine whether the transmitting or receiving end of the system is at fault. Again examining of the transmitter output is a good guide. If the transmitter is faulty, again, a logical following through of the signal waveforms from encoder to output is the only general guide that can be given. If the receiving system is suspect, the output of the I.F. amplifiers at the detector stage, should be examined. Here in particular, experience is essential, since the waveform of the signal, its size, the amount of noise permissible, varies between one receiver-decoder design, and another.

Synchronisation faults are easily detected by switching the transmitter on and off. If the synchronisation is operating incorrectly the control stick functions will jump from one servo to another. If the fault is of a different nature, say "twitching" of all servos, then examine the waveform at a channel output. If this is jumping around when the transmitter has been found to be correct, the decoder can only be at fault, if the receiver output is stable.

Faults in the receiver itself are perhaps the most difficult to find, since experience is required in knowing the correct waveforms through the circuit. Also the signal amplitude up to the first I.F. amplifier is usually too low to be examined without specilaised test equipment.

Tuning faults show up by a poor response to adjustment of the tuning cores, and here the oscilloscope waveform permits the response either side of the peak to be examined. Low sensitivity, assuming that the receiver was once correct, will either be due to a faulty tuneable component, or a transistor stage that has developed a fault. Other than an obvious mechanical failure of a component, it is often a substitution that is necessary to find the rogue component.

"Soft" radio link errors are the most difficult of all to find and here experience with the particular equipment and test procedure is essential. We have found that misleading results on the bench can be caused by such items as fluorescent lights, under floor wiring in the workshop, and even large metal objects around the room. In a manufacturers workshop, final alignment will be done under the same conditions, and as a result the exact response required of the system is known. Under conditions of weak signal, i.e. with the transmitter aerial removed, familiarity with the system tells the operator at what distance solid control should be maintained as the transmitter is moved away from the receiver. With a different workshop layout, it is quite possible that completely different results will be obtained. The person without this familiarity with the system will have to perform range tests in order to determine the exact symptoms of a fault, and then set up the equipment in the workshop and reduce the signal strength at the receiver, to simulate the fault. When the trouble has apparently been cured, a range check should then be performed under as near operating conditions as possible.

Temperature Problems

Faulty operation related to temperature is easiest to investigate if an aerosol freezer is available. The equipment can be set up under workshop test conditions, and individual components cooled until the faulty one is located. If a freezer is not available the process becomes more tedious since the equipment has to be either heated or cooled as a complete unit. With this latter method, the waveforms have to be examined for malfunction as the temperature changes.

The most common problem with temperature occurs with disc ceramic and electrolytic capacitors. With disc ceramic types the temperature coefficient can be poor with the result that tuning shifts. With electrolytic types, the problem is often mechanical in that the metallic parts used in their construction contract as the temperature falls. If the relative movement is sufficient, failures occur.

Transistors tend to loose gain as temperature falls, but a good design should have taken account of this. However, a failing encapsulation can produce faults as the temperature is lowered.

Germanium transistors used in the output of servo amplifiers can become leaky i.e. an emitter-collector current leak, as temperature rises. This fault is easily diagnosed since the servo becomes sluggish in one direction as temperature increases. Again good design will have selected a transistor that maintains the required characteristics over normal operating temperatures, and trouble will then only be experienced if the component is faulty.

CONCLUSION

In preparing this book we feel that our own understanding of digital proportional has been enhanced. It is hoped that the reader will feel the same, and that no longer will terminology appearing in various advertisements, instructions, magazine articles, etc., hold an air of mystique.

To us, digital proportional equipment appears remarkably simple considering the performance that it achieves, and we have great respect for those designers who produce the original concept. Manufacturers have taken this and implemented it in such a way that the modeller now has available a range of equipment, that if produced alongside similar electronic equipment, would cost at least three times the price.

If the equipment is treated with the respect it deserves, it should give many years of useful service. This book does not cover points concerning operation, since these are covered in the "Radio Control Guide". We would like to feel that by producing these two volumes, we have made a small contribution to the readers understanding and enjoyment of proportional equipment.

The following appendices provide constructional notes on the Microtrol system, the circuits of which were used for descriptive purposes within the book. We cannot make claims of originality for the design, since we have only taken well proven features and combined them into a single system. We have been rewarded by a reliable performance over a long period of time. Several examples have been constructed by different people, all with the same success, and we are confident that the experienced home constructor, with the necessary test equipment, will achieve the same results. However, as stated in the Introduction, this is not intended to be a constructional handbook and we would strongly advise the less experienced constructor to attempt commercially kitted equipment, where the manufacturer can offer an after sales back up for his product. We regret that we cannot enter into any correspondence concerning the Microtrol system.

APPENDIX I

MICROTROL TRANSMITTER

THE transmitter circuit is illustrated in Fig. 50, and component references in that drawing correspond to the parts list and p.c. layout in this Appendix.

The control sticks are manufactured by Skyleader Radio Control and may be purchased in kit form. If the potentiometers are not supplied with a flat on their shafts, the initial test procedure should be performed prior to assembling the stick units, in order to determine the correct position for the flat.

Assembly Notes

1. Cut p.c. board to size and drill all holes 0.75mm.
2. Open out holes where required to suit, fixing screws, preset pots., coil formers, coils, trimmer capacitors, lead tie down (holes AA), and eyelet for aerial lead.
3. Mount coil formers, with the nuts on the component side of the p.c. board. Wind coils L1, L2, L3. Start holes of windings are marked S, all coils are wound clockwise up the former, finishing at F. L3 tap point is brought out at the appropriate number of turns. L2 is wound around the outside of L1 on the former.
4. Wind L4 and L5 on a mandrel, the sense of the winding being as shown on the layout drawing. Bend leads radially to coil for mounting. Coils should be mounted so that there is 0.1inch clearance between the windings and the p.c. board.
5. Remove the adjustment screws from the trimmer capacitors and cut a screwdriver slot at the threaded end. This will allow adjustment of the trimmers from the back when the p.c. board is assembled in the case. Reassemble trimmer capacitors, ensuring that insulation pieces are in the correct position, and mount on p.c. board.
6. Mount crystal socket.
7. Mount all other components.
8. Connect all fly leads, twisting together in groups as appropriate, i.e. 3 leads to each pot., 2 leads to battery (via switch), 2 leads to output meter.
9. Tie down fly leads in holes AA located near each take-off point.
10. Connect control pots. to leads but do not mount stick units on p.c. board until initial checkout completed. The lead in each group from the positive supply rail should be connected to the right hand tag when the potentiometer is viewed from the shaft, with the tags at the top.
11. Connect battery via a switch.
12. Solder an eyelet, rim on component side, into aerial lead hole.

Initial Checkout

1. Remove crystal.
2. Set trimmer potentiometers to mid-point of carbon tracks.
3. Disconnect one end of R24.
4. Connect oscilloscope to collector of VT10 and ensure clock waveform is present, the short half of the cycle being approximately 8 milliseconds and the second half considerably longer.
5. Connect oscilloscope to collector of VT11 with input triggered off of negative edge at VT10 collector. Output should be a positive pulse which

should be adjusted for 1.5 milliseconds width by rotating the control pot. VR8, on the stick assembly.

6. Connect oscilloscope to VT12 collector, and trigger off of negative edge at VT11 collector. Adjust VR9 for 1.5 milliseconds pulse width.
7. Repeat set up procedure for each control pot. through the encoder.
8. Observe waveform at collector of VT1. This should resemble waveform of VT1 in Fig. 44.
9. Repeat for VT4.
10. Observe waveform at collectors of VT2 and VT3. This should be a train of negative pulses corresponding to the channel transitions, followed by a synchronisation pulse. Trigger oscilloscope off of collector VT10.
11. Observe that moving the control sticks changes the width of the channel command pulses over the range 1-2 milliseconds.
12. Adjust the trimpots together with the control potentiometer centring to achieve the same range of variation on each channel.
13. Observe that the synchronisation pause is of variable length.
14. Re-connect R24. Observe that the synchronisation pause is shorter, corresponding to approximately the shorter half of the clock cycle. The pause should maintain a fixed length even when the control sticks are operated.
15. Connect 10K resistors at the free end of oscilloscope test leads.
16. Insert crystal and connect an aerial.
17. Connect earth lead of oscilloscope to battery negative, and signal lead to base of aerial.
18. Adjust L1 core until R.F. output is detected, and locate the core position where the local oscillator commences to operate. Screw the core into the coil a further turn.
19. Adjust L3 core to produce maximum signal.
20. Adjust VC1 and VC2 and observe that signal level changes. It may not be possible to achieve a proper peak until after assembly into case.

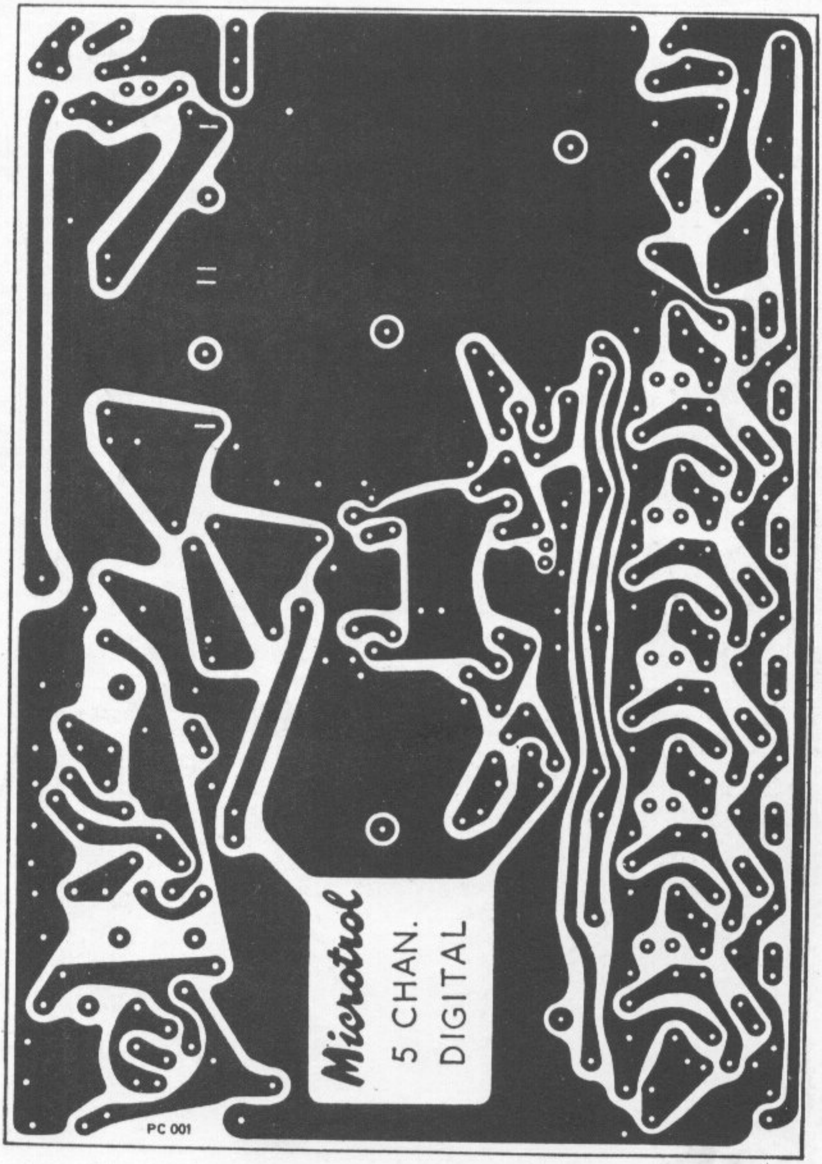
Assembly into Case

1. Construct case from 20SWG aluminium. (18SWG may be used and the top and bottom strengthening plates can then be omitted.) Epoxy strengthening plates in place, also bracket for auxilliary control pot.
2. Punch aerial socket mounting hole towards the front of the case since it must clear p.c. board.
3. Mount 9.6v, 500mAHr. battery at bottom of case, to clear p.c. board and control potentiometers. Battery may be secured by neoprene tube straps with screws threaded into the ends.
4. Fit charging socket and make an earth link between battery negative and the case.
5. Fit switch, meter and neck strap hook, and assembled stick units.
6. Wire p.c. board into case and connect switch and meter. Allow sufficient lead length to drop the p.c. board away from the stick units, but avoid excessive wire lengths.
7. The aerial lead is solid 18SWG tinned copper wire soldered to the aerial mounting screw. This lead passes through the eyelet in the p.c. board and is soldered in position after the p.c. board has been mounted on the stick units.

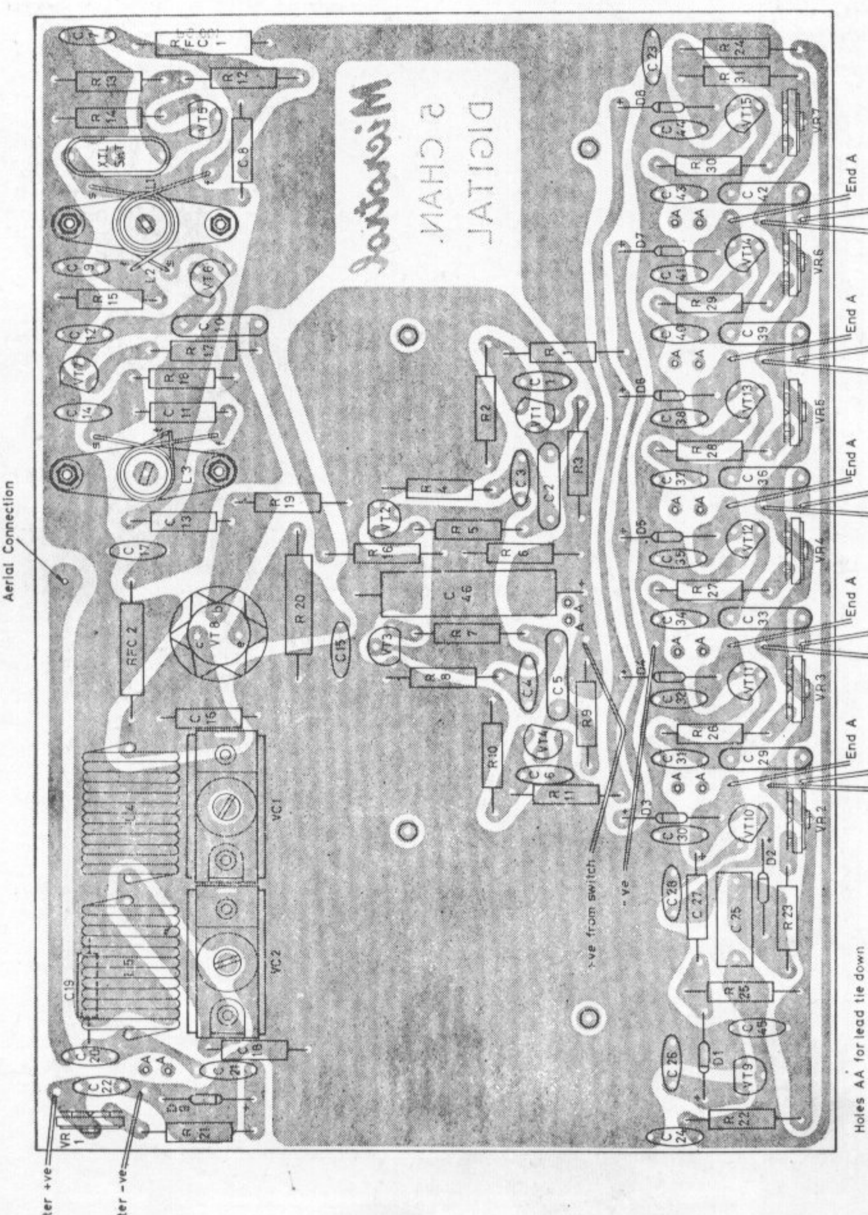
Final Alignment

1. Either use a demodulation probe on the oscilloscope or else make a 4 turn, 1cm. diameter coil and connect this between earth and signal leads of the oscilloscope.
2. Position the probe or test coil in close proximity to the aerial.
3. Adjust VR1 for maximum resistance.
4. Observe R.F. output waveform and alternatively adjust VC1 and VC2 to achieve a peak output.
5. Adjust L3 for peak output.
6. Check setting of L1 as described before.
7. Grasp Tx case firmly and fine adjust L3, VC1 and VC2.
8. Set VR1 to produce 80% deflection on meter.
9. Observe functioning of encoder by moving control sticks.

10. Seal tuning adjustments with candle wax.
11. Final test should be in conjunction with the receiver and may entail adjustment of VR2 to achieve correct synchronisation.



TRANSMITTER P.C. BOARD ACTUAL SIZE



TRANSMITTER COMPONENT PLACEMENT

Transmitter Components

R1	22K	R11	22K	R21	68K
R2	2.2K	R12	10K	R22	4.7K
R3	4.7K	R13	3.3K	R23	100K
R4	2.2K	R14	220	R24	33K
R5	2.2K	R15	100	R25	33K
R6	1K	R16	2.2K	R26	100K
R7	2.2K	R17	2.2K	R27	100K
R8	2.2K	R18	100	R28	100K
R9	4.7K	R19	100	R29	100K
R10	2.2K	R20	4.7	R30	100K
				R31	4.7K

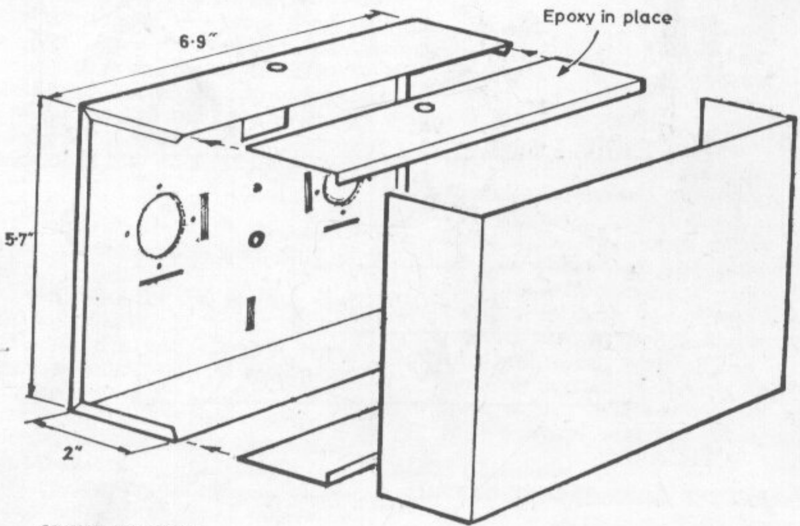
All resistors $\frac{1}{4}$ watt high stability type, $\pm 10\%$ tolerance (Radiospares $\frac{1}{4}$ watt "Hystab" or Iskra type E12) R20 may be Radiospares 1 watt wire wound.

C1	.001μF	disc ceramic. Radiospares.
C2	.047μF	polyester. Mullard C280 Series.
C3	.001μF	disc ceramic
C4	.001μF	disc ceramic
C5	.047μF	polyester
C6	.001μF	disc ceramic
C7	.047μF	12v disc ceramic. Radiospares.
C8	47pF	polystyrene. Radiospares.
C9	.01μF	18v disc ceramic. Radiospares.
C10	.01μF	polyester. Mullard C280 Series.
C11	22pF	polystyrene. Radiospares.
C12	.01μF	18v disc creamic
C13	100pF	polystyrene. Radiospares.
C14	.047μF	12v disc ceramic
C15	.047μF	12v disc ceramic
C16	47pF	polystyrene. Radiospares.
C17	.047μF	12v disc ceramic
C18	100pF	polystyrene
C19	47pF	polystyrene
C20	.01μF	100v disc ceramic or Mullard C280 Series.
C21	4.7pF	disc ceramic or Radiospares Silver Mica.
C22	.001μF	disc ceramic
C23	.047μF	12v disc ceramic
C24	.001μF	disc ceramic
C25	.15μF	polyester. Mullard C280 Series
C26	.001μF	disc ceramic
C27	1μF	25v Tubular tantalum. Econotan Type CT105

C28	.001μF	disc ceramic	C36	.047μF	polyester
C29	.047μF	polyester	C37	.001μF	disc ceramic
C30	.001μF	disc ceramic	C38	.001μF	disc ceramic
C31	.001μF	disc ceramic	C39	.047μF	polester
C32	.001μF	disc ceramic	C40	.001μF	disc ceramic
C33	.047μF	polyester	C41	.001μF	disc ceramic
C34	.001μF	disc ceramic	C42	.047μF	polyester
C35	.001μF	disc ceramic	C43	.001μF	disc ceramic

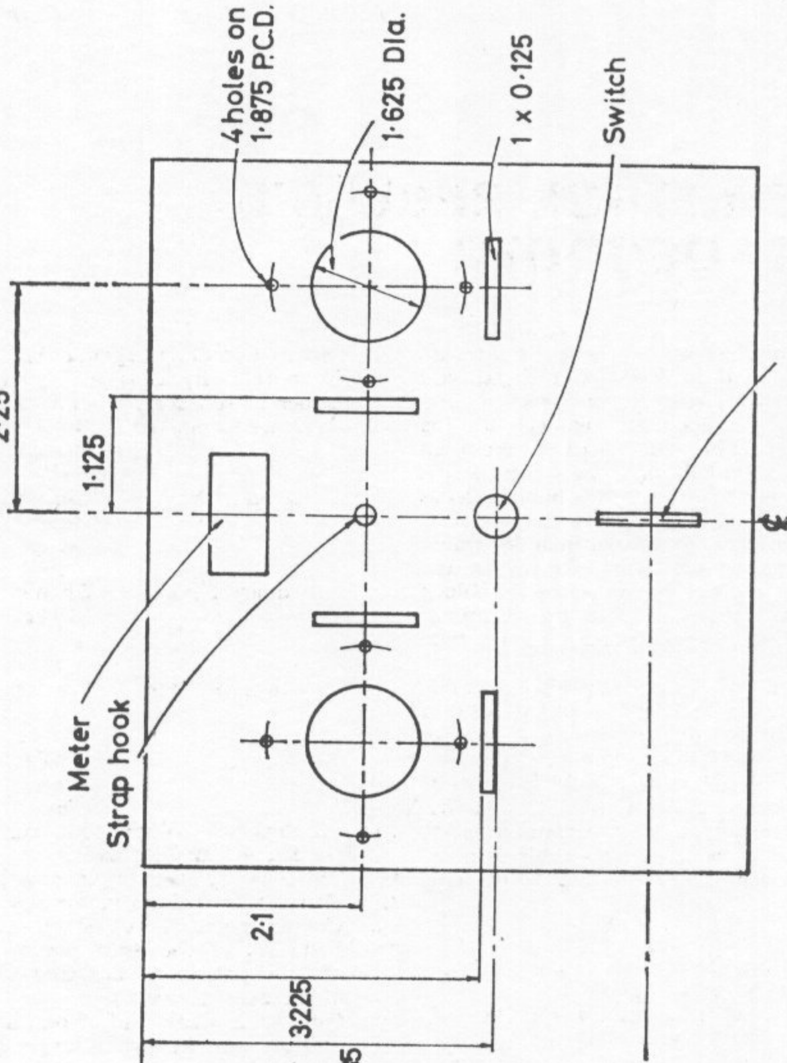
C44 .001μF disc ceramic

C45	.047μF 12v disc ceramic
C46	50μF 25v tubular electrolytic. Wima Printilyt I.
VR1—VR7	47K Egen Preset. Miniature vertical type.
VR8—VR12	5K Moulded Carbon Potentiometer, to suit control sticks.
D1—D8	1N4148 or 1N 914
D9	0A91 Mullard.
VT1—VT7	2N3706 Texas Instruments.
VT8	2N3553 Texas Instruments.
VT9—VT15	2N3706
RFC1—2	10μH Cambion moulded type or Radiospares 1 Amp T.V. Choke.
VC1—2	40pF Compression Trimmer Capacitor. Radiospares.
L1	8 turns 28S.W.G. on 6.5mm former.
L2	2 turns flex over L1
L3	3 + 9 turns 28 S.W.G. on 6.5mm former.
L4—5	12 turns 18 S.W.G. airspaced, wound on 11.5mm mandrel.
SW1	Single pole toggle switch. Must be suitable for low voltage applications.
M1	200μA full scale deflection edgewise meter.
XTL	3rd overtone. Plug in type with socket.
Aerial	140cm (55") telescopic. and socket.
Battery	9.6v 500DKZ Deac.
Charging Sockets	Radiospares miniature type, 1 each red and black.
Stick units	Skyleader or Kraft.
Auxillary control lever	S.L.M. or Horizon.



20 SWG Aluminium
Inside dimensions stated

TRANSMITTER CASE



TRANSMITTER CASE FRONT DETAIL

MICROTROL RECEIVER AND DECODER

THE receiver and decoder circuits are illustrated in Figs. 59 and 74, and component references correspond to parts lists and component layouts in this Appendix. The p.c. board is mounted in the lower half of a two part plastic case with the servo and battery leads emerging at one end, and the aerial at the other. The p.c. board includes tracks to provide junction points for the battery leads to the servo connectors, a block connector being used for three channels, and separate connectors for the other two.

A slide switch for frequency selection is mounted in the top half of the receiver case, with the toggle projecting through the case end above the cable exits. A multipole switch is used with the contacts connected in parallel in order to minimise the risk of failure and the connections to this are kept as short as possible to minimise electrical noise pickup from servo motors.

Assembly Notes

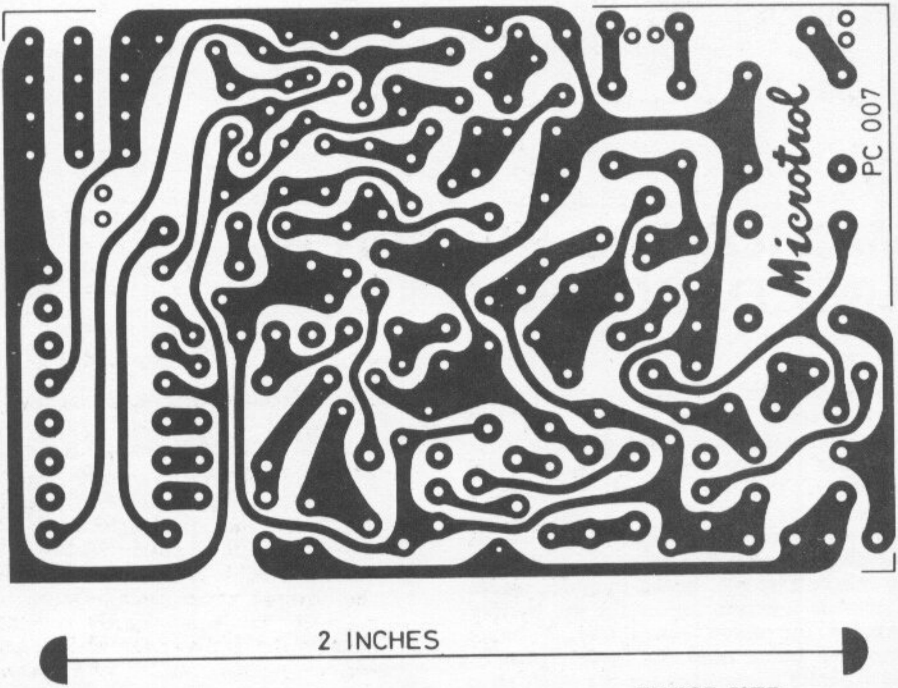
1. Cut p.c. board to size and drill all holes 0.75mm.
2. Open out holes for coil formers and I.F. transformers.
3. Wind coils L1 and L2. Fit components mounted on coil assemblies noting links on L1.
4. Mount coils, I.F.T.s and crystals.
5. Mount transistors, diodes and R.F.C.
6. Mount resistors.
7. Mount capacitors, taking care to insert tantalum electrolytics observing polarity.
8. Mount decoder integrated circuit.
9. Connect all fly leads, and group together to suit connectors as follows.

Battery Connector	+4.8v +2.4v Ov
Aileron Servo	Chan 1 +4.8v +2.4v Ov
Auxiliary Servo	Chan 5 +4.8v +2.4v Ov
Block Connector	Chan 2 Chan 3 Chan 4 +4.8v +2.4v Ov

10. Fit links to block connector for +4.8v, +2.4v Ov plus.
11. Tie down fly leads to p.c. board.
12. Connect fly leads to connectors.
13. Connect aerial and tie down.
14. Bend solder tags flush across back of multi-pole switch and connect sets of contacts in parallel.
15. Cut slot in cable end of one half of receiver case for switch toggle clearance and mount switch.
16. Connect crystals to switch, tie down leads to p.c. board and cover leads at switch end with impact adhesive.

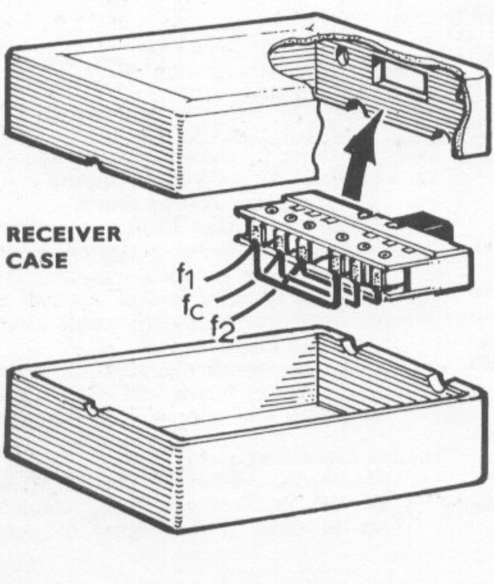
Initial Checkout

1. With 10K resistors at the free ends of the oscilloscope leads, connect up to observe the signal between



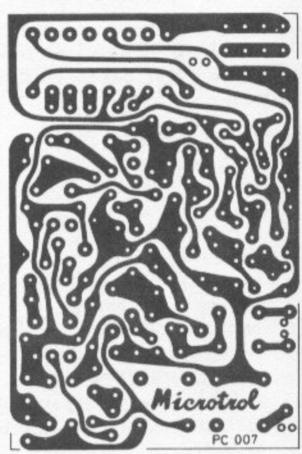
TWICE SIZE

MICROTROL RECEIVER DECODER P.C. BOARD



RECEIVER CASE

ACTUAL SIZE



VT5 collector and +4.8v.

2. Connect battery and switch on transmitter. The demodulated signal should be observed.
3. Remove the transmitter aerial and position transmitter so that a signal is still evident. Adjust L1 and L2 to achieve maximum signal amplitude.
4. Position transmitter to produce a low level signal and adjust IFT1, IFT2 and IFT3 in turn to give maximum signal.
5. Progressively move the transmitter away whilst making fine adjustments to both coils and I.F. transformers.
6. Observe the output of the auxiliary channel on the oscilloscope. Check that it follows the correct control movement and switch the transmitter on and off several times to ensure that correct synchronisation occurs. Adjust VR2 in the transmitter about $\frac{1}{8}$ turn either side of its nominal position and ensure that synchronisation is maintained when switching on and off. Adjust VR2 if necessary. Should insufficient adjustment be available, R23 in the receiver may be reduced to 6.8K.
7. Observing the auxiliary channel, ensure that only a single output pulse occurs in each frame of information, for all channel pulses at their maximum width. If more than one pulse appears, the value of R23 is too low and should be increased.

Assembly into Case

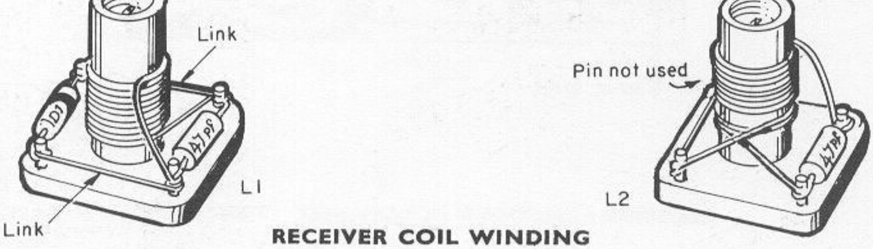
1. Fit the p.c. board into the lower half of the receiver case.
2. Route the fly leads and aerial to the cable exit holes.
3. Ensure that the slide switch in the upper case half clears all components.
4. Glue a piece of sponge rubber into the upper half of Rx case, to hold p.c. board in position when case is assembled.

Final Alignment

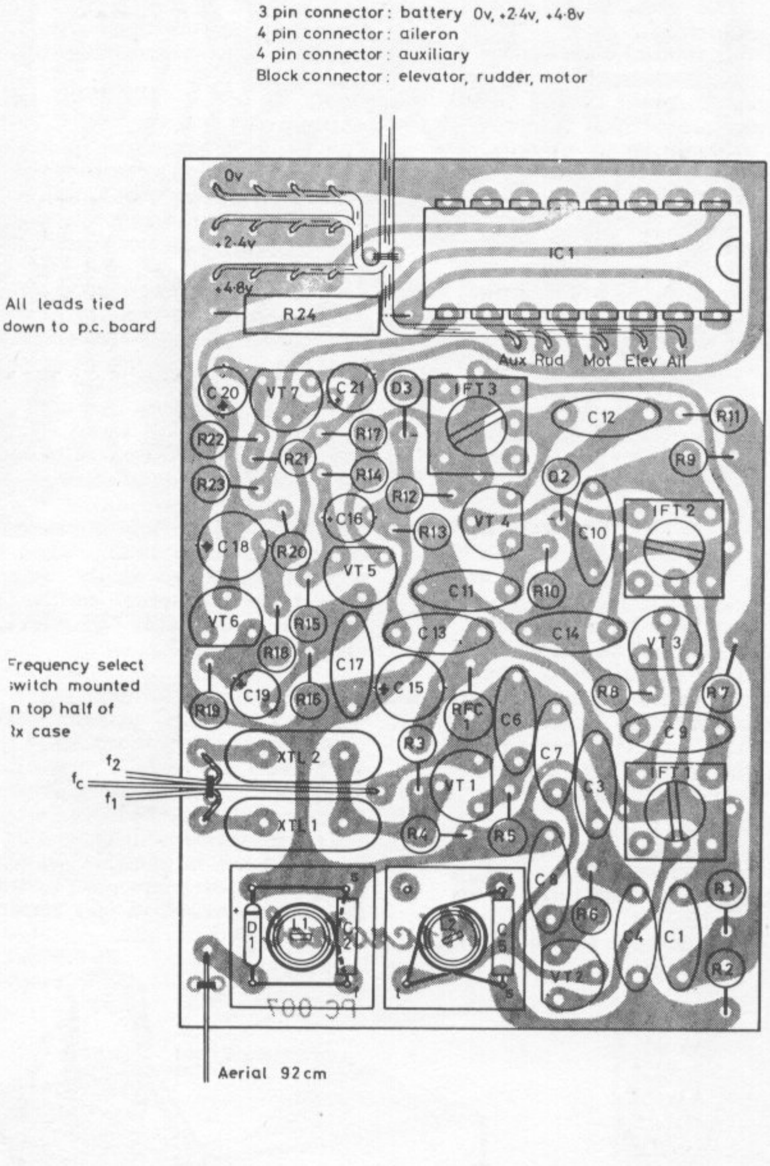
1. The receiver alignment procedure should be repeated as detailed before.
2. Seal turning coils with candle wax. The I.F. transformers should be sufficiently tight to not require sealing.
3. Perform a range test on the equipment with servos attached. This is best done with the aid of an assisttant, but should this not be possible, at extreme range, coil up receiver aerial and switch on and off. This will cause servos to travel towards one limit of throw. Uncoil the aerial and switch on, and the servos should pull back into position.
4. There should be no tendency for the equipment to swamp when the receiver aerial is lightly touched against the transmitter aerial. This can only occur with bad misalighment of L1 and L2.

Battery Harness

1. The 4.8v centre tapped battery should be securely taped up to give support to the leads.
2. A double pole switch is required to switch both 0v and +4.8v. It is not necessary to switch +2.4v.
3. It should not be possible to charge the battery with the receiver switched on otherwise damage may result to the integrated circuit.



RECEIVER COIL WINDING



RECEIVER / DECODER PLACEMENT TWICE SIZE

Receiver Components

R1	15K	R9	2.2K	R17	220
R2	2.2K	R10	470	R18	100K
R3	100K	R11	4.7K	R19	22K
R4	47K	R12	1K	R20	470
R5	1K	R13	33K	R21	10K
R6	1K	R14	47K	R22	47K
R7	2.2K	R15	100K	R23	10K
R8	470	R16	4.7K	R24	2.2K

All resistors ¼ watt high stability type ±10% tolerance (Iskra E12 or R.S. Components "0.25W Hystab.")

C1	.047µF	12v disc ceramic R.S. Components
C2	47pF	polystyrene. R.S. Components
C3	.047µF	12v disc ceramic
C4	.047µF	12v disc ceramic
C5	47pF	polystyrene
C6	10pF	disc ceramic
C7	4.7pF	disc ceramic
C8	.047µF	12v disc ceramic
C9	.047µF	12v disc ceramic
C10	.047µF	12v disc ceramic
C11	.047µF	12v disc ceramic
C12	47pF	polystyrene
C13	.047µF	12v disc ceramic
C14	.047µF	12v disc ceramic
C15	47µF	6.3v bead tantalum. R.S. Components
C16	4.7µF	35v bead tantalum. R.S. Components
C17	.047µF	12v disc ceramic
C18	47µF	6.3v bead tantalum
C19	4·7µF	35v bead tantalum
C20	2·2µF	35v bead tantalum
C21	4·7µF	35v bead tantalum

VT1—4	2N4126	Motorola.
VT5	2N3706	Texas Instruments.
VT6	2N3702	Texas Instruments.
VT7	2N3706	Texas Instruments.

D1—2	1N4148 or 1N914.
D3	OA91 Mullard.
IFT1	Mitsumi (Toko) silicon type, yellow.
IFT2	Mitsumi (Toko) silicon type, white
IFT3	Mitsumi (Toko) silicon type, black.
RFC1	5µH Painton, or 1 Amp T.V. Choke, R.S. Components.
L1	9 turns 28 S.W.G. on 5mm former
L2	8½ + 2 turns 28 S.W.G. on 5mm former
XTL1—2	3rd overtone type, 455kHz lower than Tx crystal frequencies.
IC1	SN7496N Texas Instruments.
SW1	Slide Switch. Noble 4 pole changeover type or equivalent.
Aerial	92cm (36in.) flex.
Plastic Case	1.5 × 2.3 × 0.8 inches inside dimensions SLM type PT24.

APPENDIX III

MICROTROL SERVOS

Two servos are shown for use with the Microtrol system, one using a discrete component amplifier to suit SLM FB3 mechanics, and the other an integrated circuit amplifier to suit the Skyleader SRC1 miniature servo. The circuits in Figs. 79 and 79a correspond to the assembly drawings and component lists.

Due to the small size of these amplifiers, care must be taken during assembly since removal of suspect components is difficult.

Assembly Notes

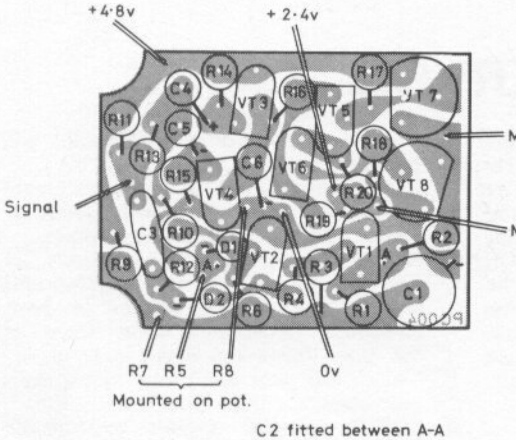
1. Cut the p.c. board to shape and ensure that it fits into the servo case.
2. Drill all holes 0.75 mm. For I.C. Amplifier only, enlarge motor case connection hole in P.C.B. to 2 mm.
3. Mount all components on p.c. board. I.C. Amplifier only, use sleeving on leads of R7.
4. Connect all leads.
5. Align pot. in servo mechanics as shown. Bend tags down, maintaining a radius to avoid fracture.
6. Discrete amplifier only. Mount R5, R7, R8 on the feedback pot., and C7, C8 on motor, using sleeving to ensure that the motor brush connections cannot touch the motor case.
7. Connect wires from feedback pot., and motor to amplifier. For the I.C. amplifier the motor case tag is soldered through the 2 mm. hole making a rigid assembly. To produce a reverse direction servo, the connections to both ends of the pot. should be reversed and also the motor brush connections.
8. Fit input leads using sleeving to give support where they emerge from the servo case.
9. Connect plug to leads.

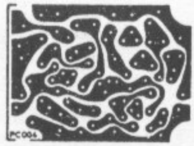
Testing

1. Connect servo to decoder output.
2. Remove motor from gear train.
3. Switch on receiver, and motor should "kick."
4. Switch on transmitter and motor will probably rotate.
5. Rotate output arm of servo to locate a position on the feedback potentiometer where motor does not rotate.
6. Rotate output arm approximately 45 degrees either side of this position and observe that motor rotates in both directions dependent upon sense of the pot. deflection about neutral.
7. Switch off and assemble motor into gear train.
8. Switch on and operate appropriate transmitter control stick. Servo should follow stick movements. If it drives hard to one end, the motor connections are reversed.
9. Should the servo response appear either sluggish, or tend to overshoot the required position, the damping control resistor should be adjusted. For the discrete component amplifier this is R2, and for the I.C. amplifier R6 provides adequate adjustment. A reduced resistor value produces a greater degree of damping.

Final Assembly

1. FB3 amplifier only. Offer up the amplifier to the gear train/motor assembly so that when assembled in the case, the suppression capacitors are partially underneath the amplifier. Place a small pieee of foam rubber on top of the amplifier so that when the servo is assembled, the amplifier is held away from the pot. solder tags. The I.C. amplifier does not need any such protection since it is correctly positioned by being mounted on the motor.
2. Slide the lower half of the servo case over the motor and amplifier, positioning lead out wires in cut out provided.
3. Gently tighten the four screws which hold the servo together. Take care not to overtighten.
4. Check servo operation to ensure that assembly has been performed correctly.

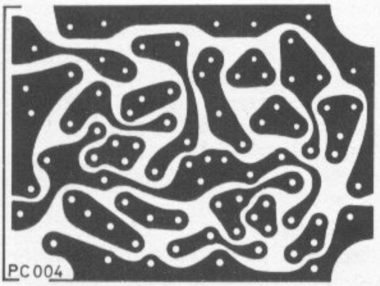




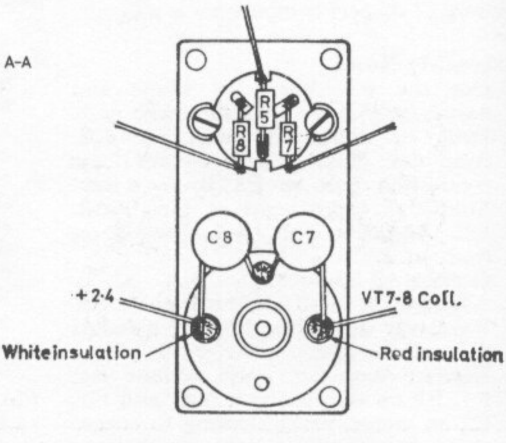
ACTUAL SIZE

SERVO AMPLIFIER PLACEMENT

TWICE SIZE



MICTROL SERVO AMPLIFIER P.C.



MOTOR AND POT WIRING

Servo Amplifier Components

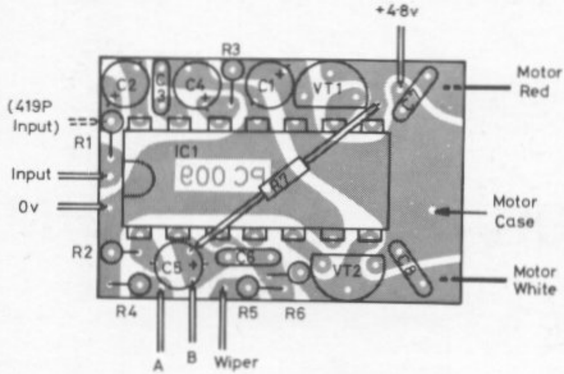
R1	100K	R8	470	R15	47K
R2	1.5M	R9	220	R16	470
R3	10K	R10	4.7K	R17	3.3K
R4	10K	R11	2.2K	R18	3.3K
R5	47K	R12	47K	R19	47K
R6	4.7K	R13	4.7K	R20	15
R7	470	R14	47K		

All resistors (except R2) Erie Type 15, ±10% tolerance
R2 Iskra type E12, ±10% tolerance.

C1	25 microfarad 6.4v Mullard Sub-miniature Electrolytic
C2	.047 microfarad polyester Mullard C280 Series
C3	.001 microfarad disc ceramic
C4	.1 microfarad 35v tubular tantalum
C5	.1 microfarad 35v tubular tantalum
C6	2.2 microfarad 15v tubular tantalum
C7	.047 microfarad 12v disc ceramic
C8	.047 microfarad 12v disc ceramic

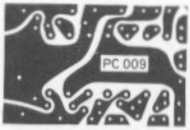
VR1 1.5K (supplied with servo mechanics)

D1—D2	1N4148 or 1N914
VT1—VT2	2N3794 Piher
VT3	2N4291 Piher
VT4—VT5	2N3794 Piher
VT6	2N4291 Piher
VT7	MPS6534 Motorola
VT8	MPS6531 Motorola
Servo Mechanics	Skyleader FB3 fitted with Mitsumi 5ohm d.c. motor.

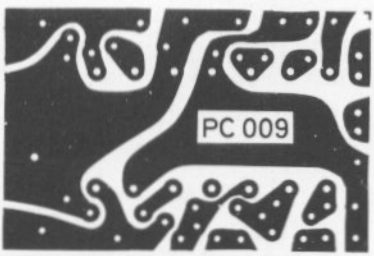


IC SERVO AMPLIFIER PLACEMENT

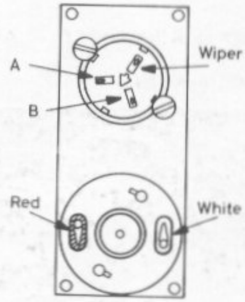
TWICE SIZE



ACTUAL SIZE



P.C. BOARD



IC Servo Amplifier Components

IC1	SRC 419 or SRC 419P	RV1	1·5K plastic film potentiometer
VT1-VT2	2N3702 Texas Instruments	C1-2	2·2 microfarad 35v bead tantalum R.S. Components
R1	10K (Not required for 419P)	C3	2200 picofarad sub-miniature plate ceramic. R.S. Components
R2	100K	C4	0·47 microfarad 35v bead tantalum. R.S. Components
R3	100K	C5	0·22 microfarad 35v bead tantalum. R.S. Components.
R4	12K	C6-8	10,000 picofarad sub-miniature plate ceramic. R.S. Components.
R5	1·2K		
R6	470K (560K for 419P)		
R7	470K (Required only for 419P)		

All resistors R.S. components 0·125W Moulded Carbon ±10% tolerance.

MICROTROL CHARGER

THE charger is designed to charge 9.6v and 4.8v 500mAHr batteries connected in series, with a charging current of 50mA. The circuit is shown on Fig. 83 and the component references correspond to parts lists and component references in this Appendix.

Assembly Notes

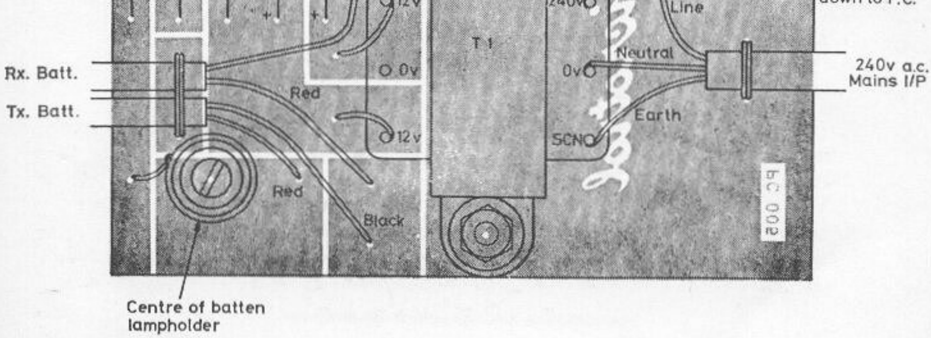
1. Construct case.
2. Cut p.c. board to shape and drill all holes 0.75mm.
3. Open out holes for transformer mounting, lamp holder and lead tie down.
4. Mount transformer, nuts to be on component side of the p.c. board.
5. Dismantle lampholder. Mount centre portion onto p.c. board not forgetting the fibre washer.

Charger Components

T1	12v sub-miniature mains transformer. Radiospares.
D1—4	1N4148 or 1N914.
R1	15 ± 10% ¼ watt high stability resistor.
L1	6v 0.06 amp round M.E.S. pilot bulb. Radiospares.
F1	250mA Cartridge fuse 20mm. Radiospares.
Lampholder	Centre portion of M.E.S. battenholder. Radiospares.
Fuseholder	Miniature Panel type 20mm. Radiospares.

Lens to fit into case, from B.M.C. trafficator lever.

MICROTROL CHARGER PLACEMENT



6. Solder fly lead to lamp holder.
7. Fit remaining components.
8. Connect batteries leads, and tie down.
9. Connect mains lead, neutral and earth to transformer, and line to fuseholder. Use sleeving to completely cover all solder tags carrying mains voltage.
10. Connect wire from fuseholder to transformer using sleeving for insulation as above.
11. Tie down mains lead.
12. Fit grommets into holes for leads.
13. Fit insulation into bottom of case.
14. Place p.c. board in position feeding leads through grommets.
15. Fit charging plugs to leads, charger positive goes to battery positive.
16. Fit mains plug.
17. Fit fuse and lamp of correct rating.

Testing

1. Connect batteries with a milliameter in series.
2. Switch on and check that charging current does not exceed 50mA. Adjust value of R1 if necessary.

CHARGER P.C. ACTUAL SIZE

CHARGER CASE DETAIL